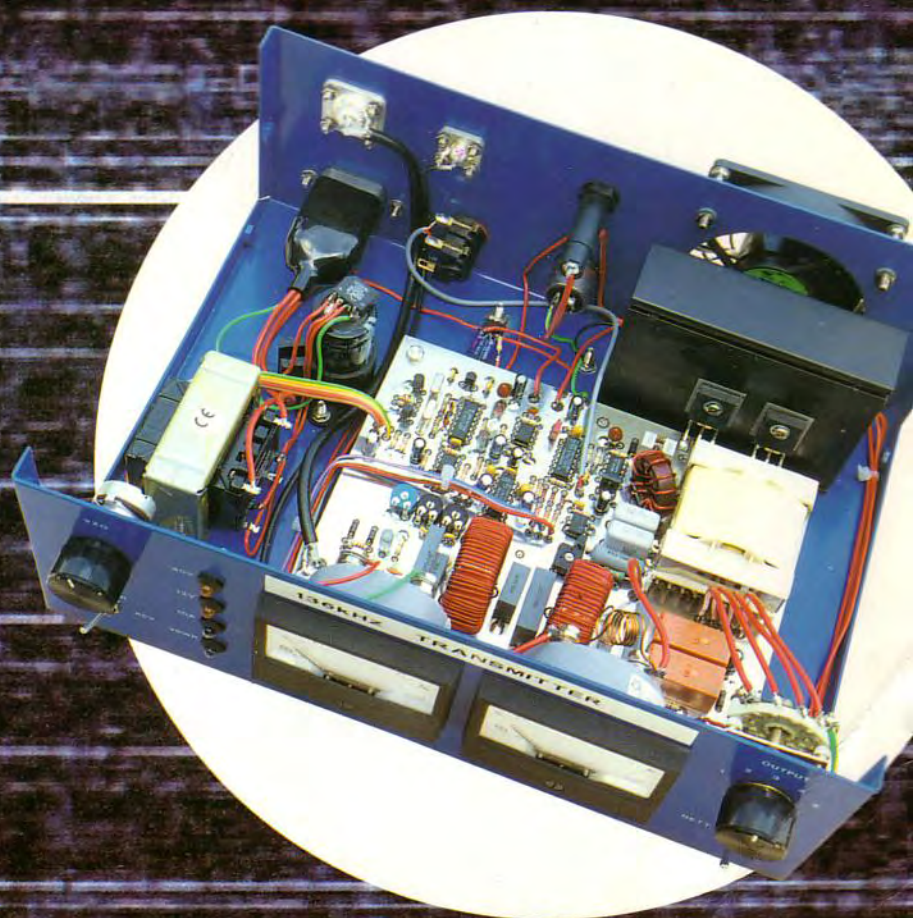


Peter Dodd
G3LDO



the
**LOW FREQUENCY
EXPERIMENTER'S**
handbook



Everything you need to transmit and receive between 50 and 500kHz

The LF Experimenter's Handbook

Edited by Peter Dodd, G3LDO



Radio Society of Great Britain

Published by the Radio Society of Great Britain
Cranborne Road, Potters Bar, Herts EN6 3JE.

First published 2000

© Radio Society of Great Britain, 2000. All rights reserved.

No part of this publication may be reproduced, stored in a retrieval system, or transmitted, in any form or by any means, electronic, mechanical, photocopying, recording or otherwise, without the prior written permission of the Radio Society of Great Britain.

ISBN 1-872309-65-8

Publisher's note

The opinions expressed in this book are those of the contributors and not necessarily those of the RSGB. While the information presented is believed to be correct, the contributors, the editor, the publisher and their agents cannot accept responsibility for consequences arising from any inaccuracies or omissions.

Cover design: Nuffield Press

Illustrations: Bob Ryan

Typography: Trevor Preece, E.P.I.C. Ltd

Production: Mike Dennison

Printed in Great Britain by Black Bear Press, Cambridge

This book is supplemented by a web page at: www.rsgb.org/books/extra/lf.htm

Preface

In the summer of 1996 the frequency band 71.6 to 74.4kHz became available to UK amateurs as an experimental band for a limited period. It was available by special permit (Notice of Variation) until June 1999 (since then the 73kHz NoV has been extended twice). Because of the lack of information regarding radio engineering techniques at these frequencies the RSGB hurriedly put together *The LF Experimenters Source Book*, extracted mainly from the LF column of *Break In* (New Zealand), *Amateur Radio*, (Australia) and *The Lowdown*, by the Longwave Club of America.

On January 30 1998 a new low frequency band of 135.7 to 137.8kHz was allocated to radio amateurs in the UK followed by an increasing number of European countries. Other countries throughout the world are expected to follow.

To meet the increasing demand for information on LF a second edition of *The LF Experimenters Source Book* was produced, which now included experimental work and constructional projects of UK amateurs

These early publications were produced quickly and the production quality was sacrificed in our attempt to make LF technical information quickly available. We have now reached the stage where we are in a position to produce a more comprehensive and higher quality book on LF. With this volume we hope we have achieved this.

The *LF Experimenters Handbook* maintains the tradition of the earlier publications, of being a collection of articles written by individuals (in a similar style to the popular *ARRL Antenna Compendium* series).

The book is divided into several sections, each covering a specific topic: *ie* receivers, transmitters, antennas, specialist modes and measurement.

Of the constructional articles, only a few have been published in the *LF Experimenters Source Book* and most, because of their experimental nature, have not been previously published at all; consequently some projects may not have the detail expected of some RSGB publications.

Most of the dimensions in this book are metric. In a couple of cases it has not been possible because formulae involving imperial units is involved.

One very interesting publication on LF is *The Lowdown*, by the Longwave Club of America. The LWCA was organised in 1974 to promote DXing and experimenting on frequencies below 530kHz and in the 1750 metre band. Details of membership of the LWCA can be obtained from 45 Wildflower Road, Levittown, PA 19057, USA.

Acknowledgements

I am indebted to the following individuals and publishers for permission to publish material used in this chapter.

All of the authors and experimenters shown in the Contents list. These individuals freely made available details of their work. Without them this book would not have been possible. Mike Dennison, G3XDV, for permission to reproduce articles from *RadCom*. Bill Oliver, for permission to reproduce material from *The Lowdown*.

Authors and/or Publications are acknowledged in the material reproduced.

Safety

Safety should be a primary consideration when erecting antenna masts. NEVER erect an antenna and mast that could possibly come in contact with electric power lines.

If you apply 800W or so of 136kHz to a high Q aerial system, the voltage on the wire will probably be over 20kV. Most insulated wire will break down and arc across to anything nearby. Keep all antenna wires well clear of everything and use good insulators.

Remember that RF burns hurt and those voltages and currents could be very dangerous!

Some of the projects in this book require a high level of constructional experience.

No guarantees of performance are implied and no responsibility for loss or damage can be accepted.

CONTENTS

Preface	3
Chapter 1, GENERAL	
The LF station at Cesare Tagliabue 1STGC	5
LF operation from the Puckeridge Decca Station	9
Portable operation from Amberley Industrial Museum	11
Photo gallery	12
Chapter 2, RECEIVERS	
LF receive systems	14
LF converter from 136kHz to 14.136MHz	17
Receiver amplifier and loop antenna for 136kHz: Andy Guye F6CNI	19
A 136kHz high performance receiver converter: John Moore G4GVC	20
A simple signal canceller for 136kHz: Derek Atter G3GRO	20
The GW4ALG noise canceller: Steve Rawlings GW4ALG	22
Outline design for a narrow bandwidth LF receiver: Andy Talbot G4JNT	24
An audio clipper for headphones: Dick Rollema PA0SE	24
A sensitive loop antenna for 136kHz: Tony Preedy G3LNP	27
Chapter 3, TRANSMITTERS	
Crystal mixer exciters for 136kHz: Peter Dodd G3LDO	31
A transmitter for the 137kHz band: HaJo Brandt DJ1ZB	32
A crystal/mixer for 137kHz transmitter: Peter Schnoor DF3LP	34
A transmitter for 136kHz: Dick Rollema PA0SE	35
A 400 watt transmitter for 136kHz: David Bowman G0MRF	37
A 136kHz 1kW transmitter: Dave Pick G3YXM	40
A 400 watt power amplifier using 572Bs: Steve Rawlings GW4ALG	45
LF power amplifiers using a bridge configuration: Peter Dodd G3LDO	46
Pulse width modulator for class D power amplifiers: Andy Talbot G4JNT	48
Chapter 4, LF ANTENNAS AND COILS	
LF vertical antennas: Bob Vernal ZL2CA	51
A guide to LF Antennas: Peter Dodd G3LDO	54
An inverted L/V antenna for LF: Peter Dodd G3LDO	56
A helical antenna for 136kHz: Tony Baertschi HB9ASB	58
Balloon-extended vertical LF Antenna: Steve Rawlings GW4ALG	59
A limited space 136kHz antenna: Rik Strobbe ON7YD	60
The loading coil: Peter Dodd G3LDO	61
Low frequency coil Q: Bill Bowers	65
LF variometer design: Tony Preedy G3LNP	66
Chapter 5, SPECIALIST TECHNIQUES	
Narrow band techniques: Rik Strobbe ON7YD	68
Receiving and transmitting slow CW: Peter Dodd G3LDO	72
Low speed Hellschreiber on LF: Andy Talbot G4JNT	74
PSK31: a new radio-teletype mode: Peter Martinez G3PLX	75
Chapter 6, TEST EQUIPMENT	
Simple test equipment for 136kHz: Peter Dodd G3LDO	81
The ScopeMatch tuning aid: Jim Moritz M0BMU and Lech Laszkiewicz G3KAU	82
A field strength meter for the 137kHz band: Dick Rollema PA0SE	83
Measurement of antenna system impedance: Peter Dodd G3LDO	86
APPENDICES	
The earthing resistance of antennas: Dr Eng A Meissner	89
General information	95
Crossing the Atlantic on LF	97
Data on the HIP4080 full bridge FET driver	99

1

General

The 137kHz Station at 15TGC

By Cesare Tagliabue 15TGC

Central to the design of my station is a transceiver with a transmitter output on 137kHz. Originally I used a home constructed transceiver built in 1991. This transceiver is a conventional solid-state double conversion set providing general coverage reception from 0 up to 32MHz. Other characteristics are: noise floor at -126dBm with 1800Hz bandwidth, CW filter 500Hz, a dynamic range over 95dB and an S-meter calibrated in dBm as well as in S-points and microvolts.

chain and the antenna. A block diagram of the station equipment is shown in Fig 1.1.

On receive the signal from the antenna is fed, via a change-over relay in the PA to a combining circuit, which is adjusted for a ratio of signals from the main antenna and a loop antenna. With appropriate signal level adjustment and loop bearing, the received signal can be adjusted for optimum signal to noise ratio.

Details of various circuits now follow.

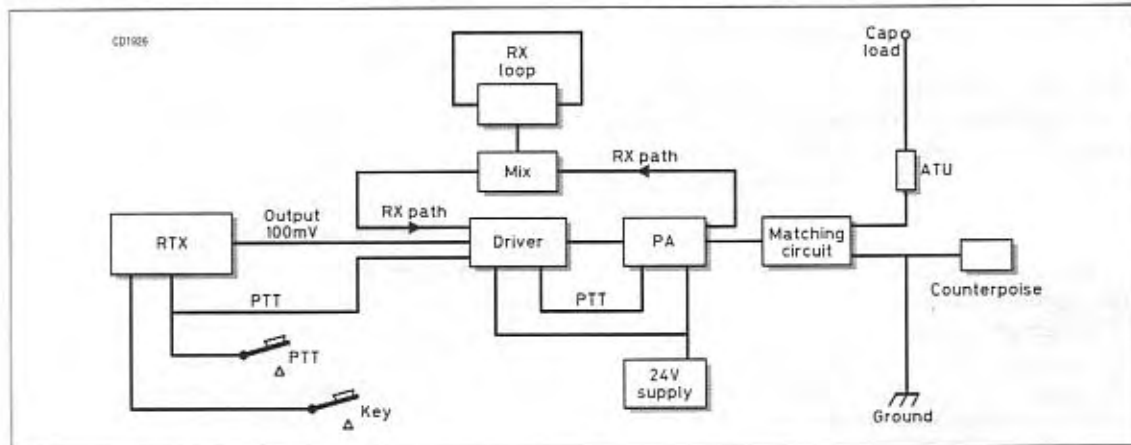


Fig 1.1: A block diagram of the station equipment used at 15TGC.

On transmit it operates on all amateur bands from 1.8 to 28MHz, with an output power of about 3 watts. In the general coverage mode the transceiver can generate a signal of about 100mV RMS at the antenna connector when switched to transmit and this has been used as an exciter for a 137kHz transmitter.

To complete the station there is a transmitter drive

Receiver/Exciter

Although the transceiver described above has adequate frequency stability for normal SSB and CW operation it is not good enough for QRSS, where the long-term stability must be kept within a few Hz.

As a consequence, a receiver/driver was constructed

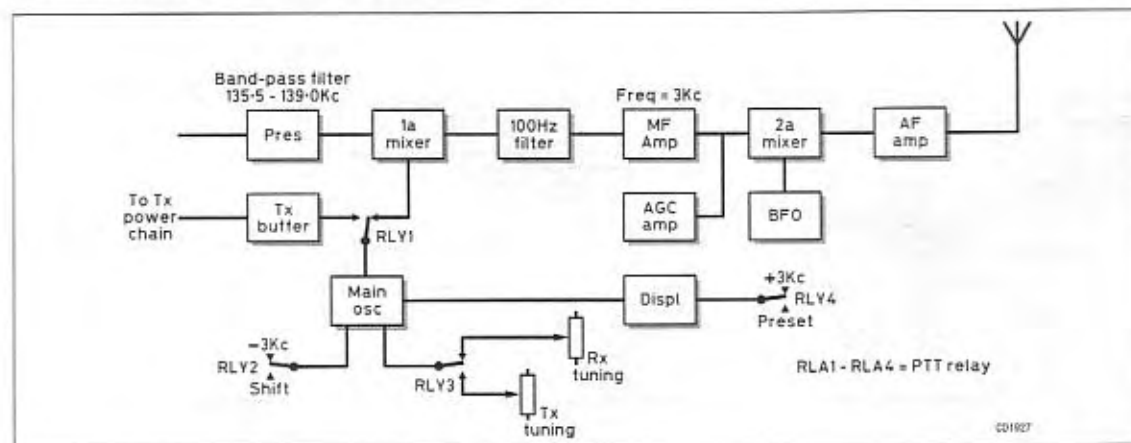


Fig 1.2: Block diagram of the 137kHz receiver/driver. The oscillator is used for transmit and receive; shifted -3kHz on receive to compensate for the +3kHz IF conversion. The display is preset +3kHz to show the correct receive frequency. In addition the receiver and transmit frequencies can be adjusted independently. The image is rejected with a bandpass filter ahead of the first mixer.

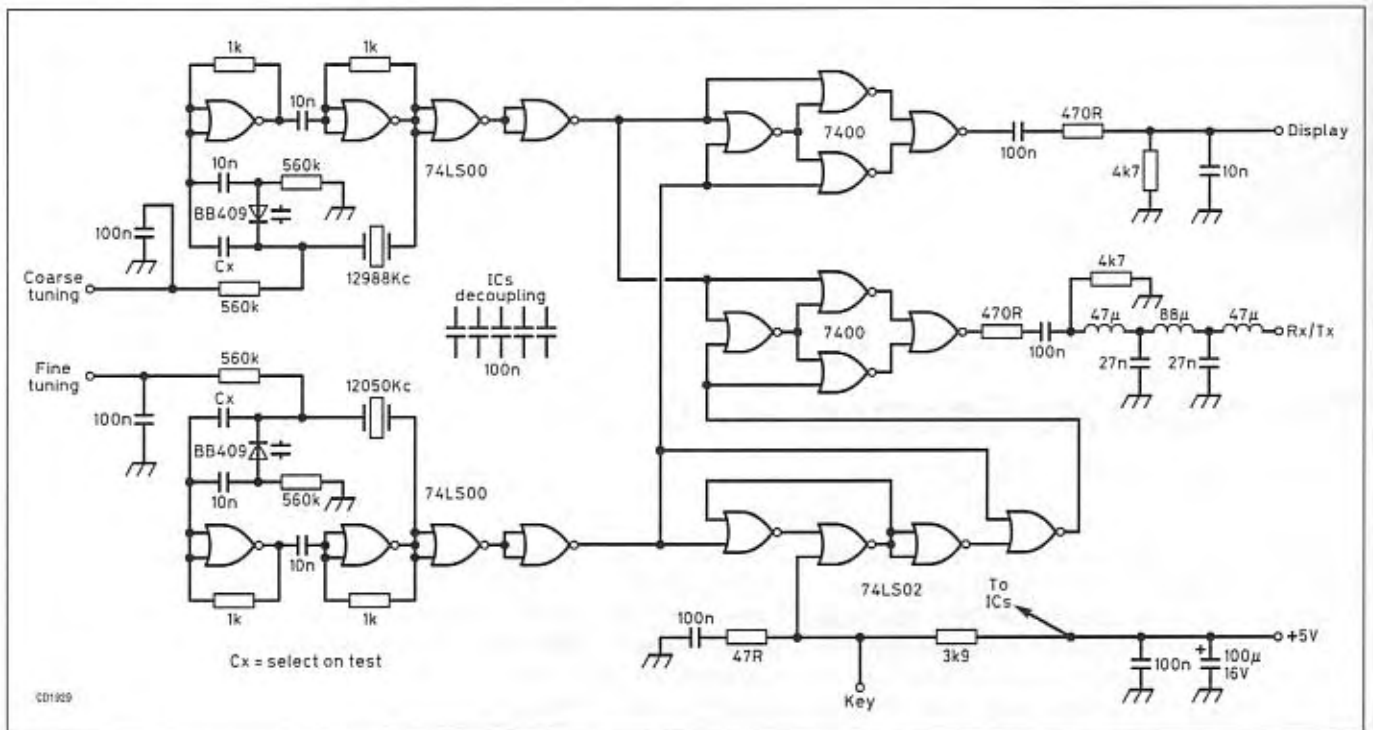


Fig 1.3: Crystal/mixer local oscillator circuit diagram. Capacitors Cx are adjusted on test to give the correct output frequency and frequency range.

specifically for the 136kHz band, the actual frequency range being 135.5 to 139kHz. This unit has sufficient stability to operate in QRSS mode and the transmit output is about 100mV RMS, which allows the use of the existing transmit circuits.

The receiver circuit can be regarded as a superhet receiver with a very low frequency IF. The choice of 3kHz for this stage makes it possible to design a simple but very effective 100Hz filter. The signal is then amplified and converted to a lower frequency using a mixer and a BFO in the conventional manner.

A block diagram of the receiver/exciter is shown in Fig 1.2.

This transceiver uses some of the modules from a synchrodyne receiver, which I originally built in 1975. Some of the original circuits, such as the mixer, the IF amplifier with its optical AGC, the power supply unit and the display with its counter and the case are used without modification. These circuits are more or less conventional. The first mixer uses an SN76514 in a standard layout.

The IF amplifier uses an SN72558 arranged as two stage variable gain amplifier, with a photo-resistor inserted in each loop and controlled by a LED fed by the AGC current. The second mixer uses the LM1596 as demodulator and the NE555 as BFO generator. The AF power amplifier uses the TAA 611 in a standard layout.

The VFO, receiver front-end and the audio filter have been designed specifically for the 137kHz and these will now be described.

Local Oscillator

The local oscillator, which also serves as carrier generator, is constructed using logic chips and is shown in Fig 1.3. It uses two 13MHz VXOs, each tuned with varicap diodes. This arrangement gives a degree of temperature compensation. The construction of the oscillators is very similar, hence any change in temperature will effect both VXOs to the same degree so any change is cancelled out.

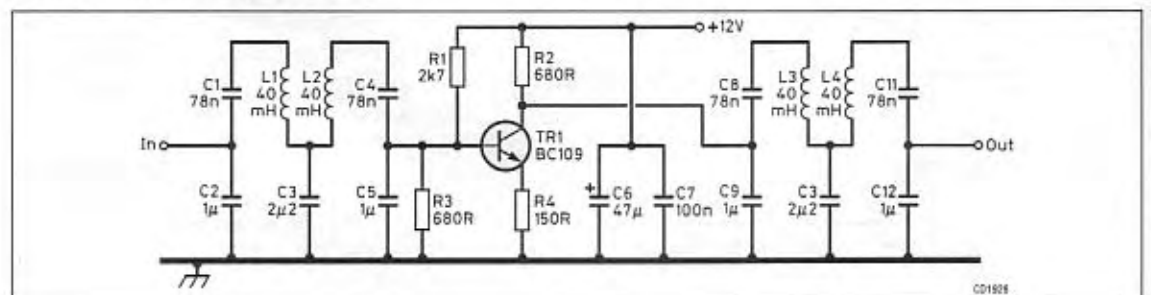
The oscillator is used for transmit and receive, shifted -3kHz on receive to compensate for the +3kHz IF conversion. The 3kHz shift voltage is input to the 'Coarse Tuning' input. In addition the receiver and transmit frequencies can be adjusted independently on transmit and receive and the adjustment voltage is applied to the 'Fine Tuning' input.

The outputs of these oscillators are then mixed and filtered.

IF Filter

As already mentioned, the choice of 3kHz as an IF frequency makes the design of a very effective filter very simple. The filter has a bandwidth of 100Hz and this is

Fig 1.4: 3kHz IF 100Hz band-pass filter, circuit diagram. R1 = 2k7; R2, R3 = 680R; R4 = 150R; C1, C4, C8, C11 = nominally 78p polyester (actual value selected on test for the correct frequency response); C2, C5, C9, C12 = 1μ polyester; C3, C10 = 2.2μ polyester; C6 = 47μ 160V electrolytic; C7 = 100n ceramic; L1, L2, L3, L4 = ferrite pot core 40mH; Q1 = BC109 or equivalent.



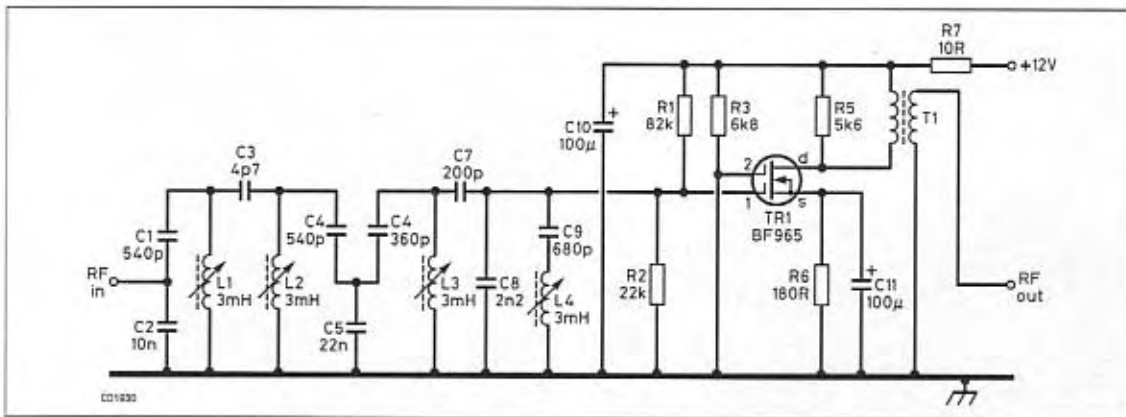


Fig 1.5: Receiver input filter and amplifier, circuit diagram.
 C11 = 100µ 16V;
 T1 = ferrite toroid 13/13;
 L1, L2, L3, L4 = 3mH slug core.

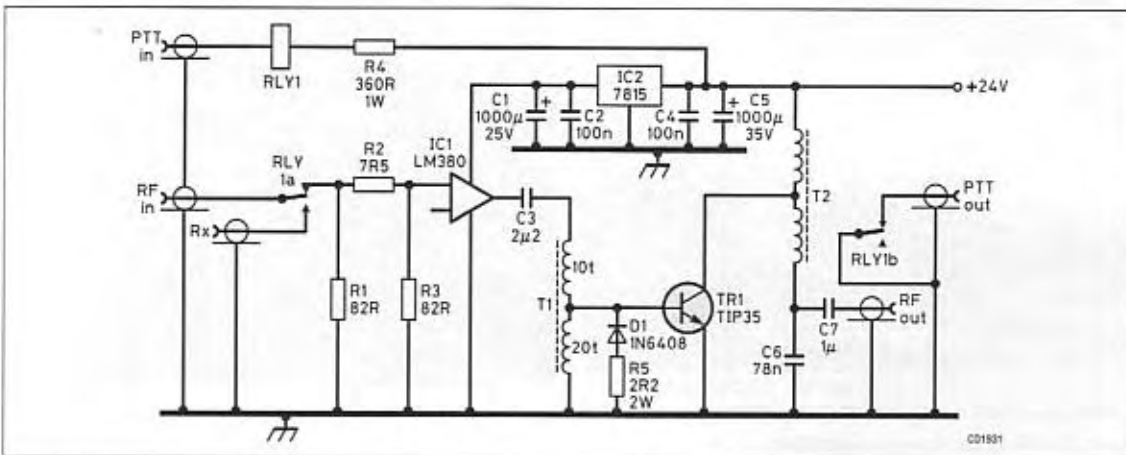


Fig 1.6: Transmitter pre-amplifier circuit diagram.
 T1 = Ferrite toroid 20+10t;
 T2 = Ferrite toroid 20+20t;
 RLY1 = relay 12V 250 ohms.

achieved using four 40mH pot cores and a few capacitors. The circuit is shown in Fig 1.4. The filter is placed ahead of the IF amplifier as shown in Fig 1.2.

Receiver Input Filter and Amplifier

The first mixer is preceded by a four-section band-pass and image rejection filter, plus a preamplifier, as shown in Fig 1.5.

This circuit provides the necessary selectivity to give adequate image rejection and gain to give the receiver the necessary sensitivity to receive weak signals at LF.

Transmitter Pre-amplifier and PA

The output from the oscillator is only around 100mV RMS so the required gain of the transmitter amplifier circuits is high. To maintain stability the amplifier is divided into two stages; a pre-amplifier, and a power amplifier.

The preamplifier, shown in Fig 1.6, uses an LM380 followed by a TIP 35 in class C. The output is about 7 watts and although the waveshape is rather distorted it is

suitable for driving the PA working almost in class D.

The PA uses four TIP 34Cs in parallel. The output is followed by an L section filter with a Q of 7, which causes the output to be sinusoidal and the harmonics reduced to an acceptable level.

The circuit of the power amplifier is shown in Fig 1.7. The choke L1 is air cored, 21 turns, 33mm winding diameter, supported by three thin fibreglass spacers, 2mm diameter silver-plated copper wire spaced 2.54mm. The coil L2 is also air cored, 14 turns, 60mm winding diameter, supported by three thin fibreglass spacers, 3mm diameter silver-plated copper wire spaced 5.08mm. The transmitter gives an output of 200W for an input of 300W DC, an efficiency of about 66%.

Fig 1.7: Transmitter power amplifier, circuit diagram.

R1 = 220R 2W; R2 = 7R5 4W; R3 = 1R 2W; R4, R5, R6, R7 = 0.1R 7W; R8 = 25R 4W; R9 = 33K 2W; C1 = 1000µ 25V; C2 = 100n ceramic; C3 = 220n polyester; C4 = 10n 250V ceramic; C5 = 18x10n mica 600V; C6 = 8x1µ polyester; C7 = 5000µ 50V; T1 = ferrite toroid 60/8t; L1 = choke 7µH see text; L2 = tank coil, see text; RLY1 = relay 12V 250ohm 10A; TR1, TR2, TR3, TR4 = TIP34C.

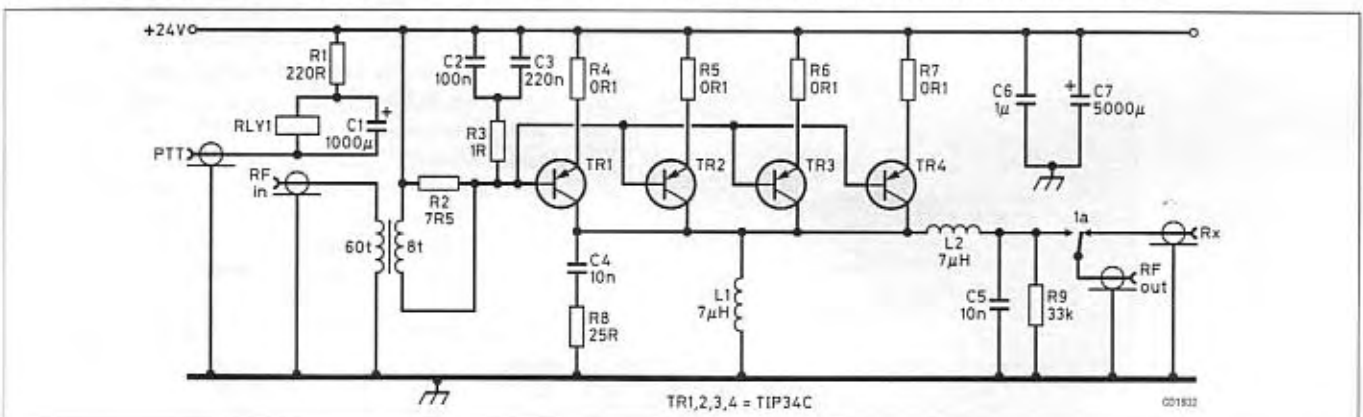
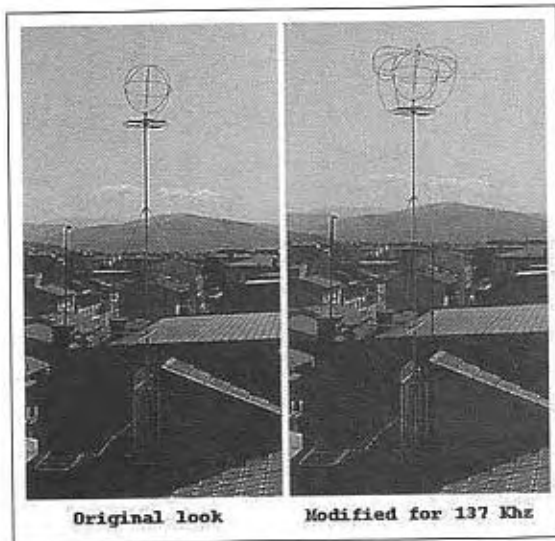


Fig 1.8: The original 160m antenna, before and after modification for 137kHz.



The supply unit is a 24V brute-force design, roughly regulated by an inductive filter with swinging choke, 76,000µ capacitor and 30Ω 30W bleeder resistor.

Antenna

The 15TGC station situated in a building just inside the town of Florence, with no space to arrange horizontal wires as loading capacitance for the antenna. The only option was to adapt the short vertical dipole built in 1991 for 160m, shown in Fig 1.8. This dipole has proved to be very effective since it was constructed. It has enabled me to make many 160m DX contacts with USA stations without difficulty, using only 100 watts power.

To get this antenna to work on 137kHz was much more of a challenge. The final arrangement is shown in Fig 1.9, which also shows the measured RF current flowing in the various parts with 200 watts of applied power.

The main changes to the original antenna include increasing the top capacitance and bonding the various nearby metallic structures. In addition many wires have been arranged over the roof underneath the antenna and connected to the ground lead to reduce losses in the walls.

With such a small antenna the losses are fairly high. It is difficult to make a theoretical analysis of the antenna itself because the layout is so unconventional.

Fig 1.9: The layout of the 136kHz antenna at 15TGC.

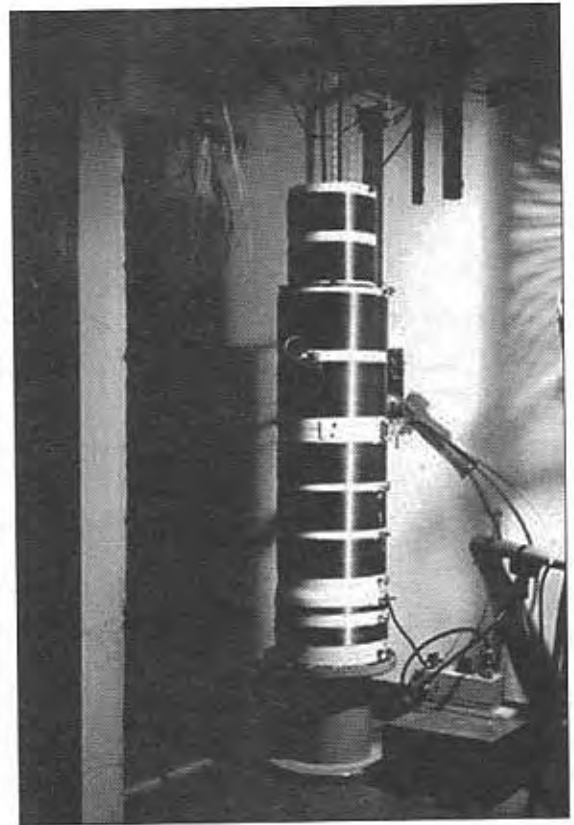
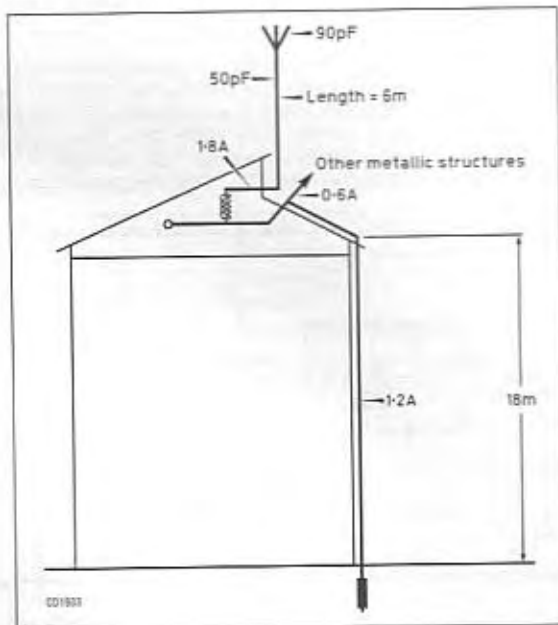


Fig 1.10: The remote controlled loading coil/variometer

Nevertheless an estimate of gain gave a figure of around -38dBi; this gives an ERP of 20milliwatts with transmitter output of 200 watts. The calculated gain turned out to be very close to a series of indirect measurements made in tests with Valerio IK5ZPV, who lives in Pistoia, 35km from Florence.

The tuning coil with its remote controlled variometer, is shown in Fig 1.10 and has a total inductance of about 9.5mH. The series resistance is around 30 ohms, therefore the Q of the coil is near 275.

The total resistance at resonance of the antenna is around 60 ohms in dry weather owing to the very good ground system of the building. However, in wet weather this value can go as high as 120 ohms or more making the antenna almost useless. This is not helped by the PVC tubing, which is used as support and may be partially hygroscopic. It is planned in the future to replace the PVC with MOPLIN rods, and upgrade the overall insulation to avoid this restriction.

For receiving, also a square loop 1.25 x 1.25m, using 75Ω TV cable, is used. This loop, with 18 turns primary winding and 1 turn secondary winding, equipped with a FET amplifier, shows a loss of about 45dBi. Its output is combined with the output of the main antenna (connections shown in Fig 1.1) resulting in a cardioid pattern useful in some cases to reduce the QRN. This loop is now located indoors; in the future the plan is to build another weatherproof loop, to be located outdoors, so being less prone to the local noise.

LF Operation from the Puckeridge Decca Station

By Derek Atter, G3GRO. Photos by Mike Dennison G3XDV

This is a description of the operation on 11 April and 14/15/16 April 2000, from the ex-Decca station at Puckeridge, near Cambridge. It was arranged to coincide with the decommissioning of the Decca station when the original equipment would be removed but mains power would still be available.

At very short notice, word was spread around all of the LF operators living within reasonable reach of the site so that one or two groups of operators could take part. All the amateur LF equipment was located in a small building at the foot of the mast. This room not only had power but it was electrically heated, which proved a boon in the cold wet days that occurred during the operation. The only original items of equipment used were the Decca mast and an RF thermocouple ammeter.

The mast was a self-supporting lattice structure, 100 metres high and supported on four massive ceramic insulators. It was fed with a copper pipe lead-in, which went to the original loading coils.

Equipment Used

For 136kHz operation, a relatively small variometer (about 500 μ H max) and a tapped toroidal autotransformer was used. One end was connected to the copper pipe lead-in and the other to the ground system.

For 73kHz operation an additional loading coil was designed and built by Lech, G3KAU, after an exploratory visit earlier in the week.

Three transmitters were employed at various times:

- The G3GRO 300W much modified BK Electronics linear amplifier used for the 1W ERP tests on 73/136kHz earlier in the week, and on Friday and Saturday for running between 1 and 5W ERP.
- The well-tried G0MRF 500W set-up.
- The G3YXM 1kW class-D rig used previously on many /P expeditions.

The receive system comprised a home brew up-converter with an input bandpass filter of about 3kHz bandwidth on both 73 and 136kHz, followed by a Mini-circuits MAR6 preamp and MC1496 mixer IC to a 10MHz IF feeding both an FT-990 and an IC-756 transceiver for most of the time. The FT-990 and converter stood up remarkably well to the very large antenna input with no sign of cross-modulation. We had a switched attenuator at the input to the converter but for most of the time it was switched out. That attenuator was used later for some comparative tests with a small 'back-garden' type antenna.

Operation

We had a few problems to start with. The first of these was getting the variometer to handle 500W RF. There were pyrotechnics and sparks resulting in VSWR trips until we realised that the capacitive voltage divider in the forward/reflected power meter in the variometer was breaking down. The problem was cured simply by bypassing the faulty SWR meter. Fortunately we had another SWR meter in line so we could keep an eye on what was going on. We also found that the ScopeMatch (see Chapter 6) brought along by Jim, M0BMU, was very



useful in setting up the matching and tuning in conjunction with an oscilloscope.

We had two operator positions side by side, one handling the 136 and 73 traffic and the second operator monitoring 136/73kHz in parallel but also handling the HF cross-band input mainly from 7.0MHz.

We found that the FSD of our RF ammeter was insuffi-

Fig 1.11: Dave, G3YXM, operating. Note the life-sustaining box of chocolate biscuits on top of the rig.

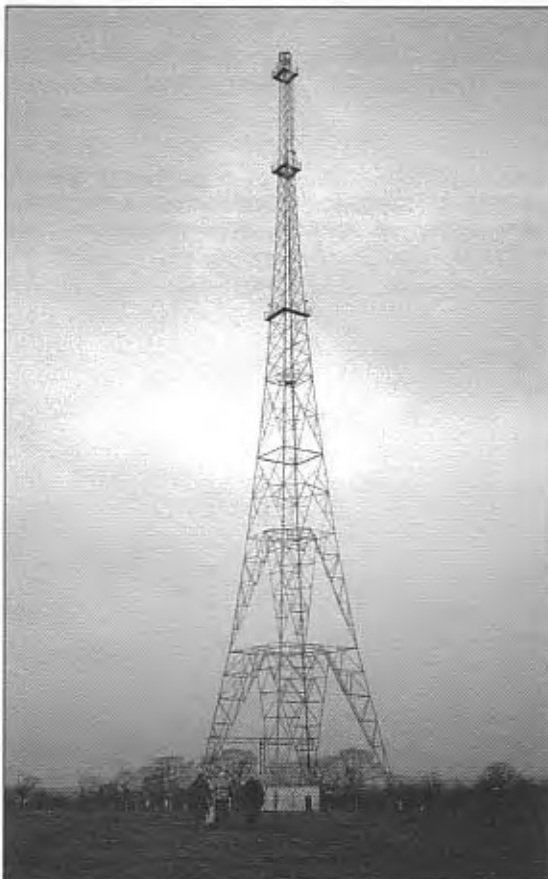


Fig 1.12: A 100-metre high Decca mast. This mast is located near Lewes in Sussex but is the same as the one at Puckeridge.

Fig 1.13: The entire mast is insulated from ground. Note the spark gap across the insulator.

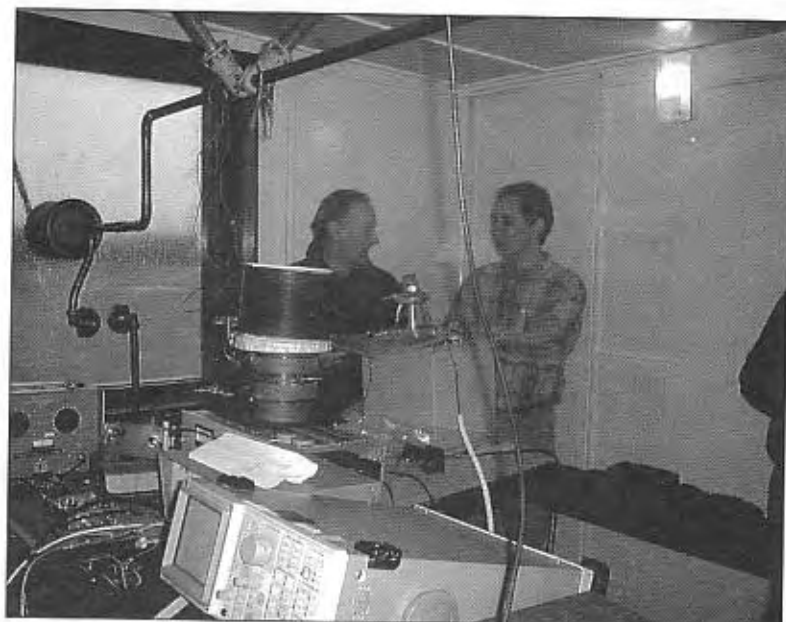


Fig 1.14: The copper pipe antenna feed conductor with the 136kHz and 73kHz G3KAU loading coils attached. Note the window insulator and the spark gap.

Fig 1.15: Some of the team on Saturday 15 April (left to right): G3XTZ, G3YSX, G3KAU, G3GRO, G3YXM, G4GVC, G4TSH, G3XDV, M0BMU (and behind the camera G4JAI). The awful weather can be seen from the quality of this picture - and it was the best part of the day.



cient for the high power operation and rescued one of the original very large 30A RF ammeters from the pile of redundant scrap units outside in the rain and pressed it into service. During the high power beacon sessions the current reading was 14A into the base of the mast, which is 5 ohms resistive in series with 3750pF! That represents an RF power of 1kW into the antenna system or 100W ERP allowing for an overall antenna efficiency of 10% on 136kHz. These beacon sessions occurred overnight on Friday (ERP was 50W) and overnight Saturday (100W ERP). Note that a special high power licence had been obtained in advance of the tests.

Summary of Results

We had about 65 contacts in total including those during the initial setting up period on Tuesday 11 April. Two-way contacts were made on both bands, and cross-band contacts from 136kHz to 73kHz and to 7.0MHz.

The best DX was a cross-band contact with Alex, UB5WF, who was transmitting on 7.0MHz and gave us RST429. His location is about 200km north of Odessa on the Black Sea (Loc KN58JQ, a distance from us of 2225km). This took place over a daylight path at 1232 UTC. It is not known what receive antenna Alex had for 136kHz but it is highly unlikely that it was a dedicated LF antenna since there was no known LF activity in Russia.

Other long distance contacts were to Valerio, IK5ZPV, (2-way on 136) who gave us RST589, IK7HSS cross band to 7.0MHz, and Neils OZ8NJ (2-way on 136).

We also got an RST 599 on 136kHz from SM6PXJ, OZ5N and HB2ASB. Two QSOs were also made on both 136 and 73kHz to EI0CF and GI3PDN. Reports on 73kHz were about two S-points down on 136kHz with the Irish stations. We also worked GJ4CBQ and GU3SQX cross-band 136kHz/7.0MHz, which was pleasing since the Loran transmission from Lessay makes it normally difficult to hear stations on 136kHz. Perhaps one of the most unusual QSOs was with Graham G3XTZ/M operating mobile on 136kHz CW while driving to the site to have a spell of operating. We also had a report from F5MAF in Toulouse, who was hearing us at 599+ on a 2 metre diameter loop at a distance of 900km and was bemoaning the fact there was no LF activity in his neck of the woods.

Many thanks are due to all those who took part, despite the very wet and cold weather, and also to those who took the trouble to give us reports. At one point on Saturday, as the shifts changed over, there were 12 people in the ATU shack at the base of the mast which represented a large slice of the active UK LF operators! They were: G3KAU, G3XDV, M0BMU, G0MRF, G3YXM, G3XTZ, G3YSX, G4GVC, G4JAI (XYL GVC), G4TSH, G3LHZ and G3GRO.

Operation from the Amberley Industrial Museum

By Peter Dodd G3LDO

The museum has a standard HF amateur station GB2CPM, which forms part of a permanent exhibition of commercial, military and amateur radio equipment. The LF experimental station is a temporary additional installation.

The location is about 7km north of Arundel on the south coast of England, and the site has been used for hundreds of years as a chalk quarry with kilns for making lime. The result is that there are small ravines dug out of a chalk hill at this location in the South Downs. The museum was originally called the Chalk Pits Museum (hence the callsign).

I was given permission to set up a station for 136kHz. This could not be located at the main station because it was impractical, public safety being a consideration.

The LF antenna is an inverted-L with a sloping 30m vertical section and a 150m horizontal section, located in a 300m wide ravine created by earlier quarrying activity.

The only solution for placing the radio equipment in a suitable location relative to the antenna is to use a four-wheel drive vehicle as shown in Fig 1.16. The loading coil, wound on white plastic fencing material, is shown to the right mounted on top of an upturned plastic dustbin. On early receive tests, the Loran signal blasted through, even though the bulk of the hill blanks the signal path to the south, which means that chalk may be a good insulator – even when wet.

The physical conditions at the site are a bit rough during the LF season, with very deep ruts filled with chalk mud and water, sometimes frozen. However, I do have access to power from the nearest building via a 150m long extension (hyper) lead.

An operating desk in the vehicle is made by folding down the back seat and simply placing a piece of plywood over it as shown in Fig 1.17. The radio equipment comprises a TS-850 for receiving and a BK Electronics audio amplifier based transmitter.

So far the main problem is that the antenna system impedance is way outside anything I have come across before. It might be due to the chalk ground but to date the problem has not been resolved satisfactorily. I have had moderate success using long radials. The interesting thing about this antenna system is that it is 'floating' above earth. If I attempt to ground it to the vehicle or mains earth the current falls to zero. My feeling is that the matching system I am using does not have the impedance range to match an antenna system that is very much different to that used at the home QTH. The floating anten-



Fig 1.16: Location of the 'shack' at the bottom of the small ravine. The secondary loading coil is on an upturned plastic dustbin and the bottom of the antenna can be seen because of the cloth markers (to prevent the tractor driver getting tangled up in it).

na system works quite well but the loading coil appears not to be in the centre of the system. An oddity of this antenna system is that the loading improves after I have been operating for some time. Obviously more research is required.

Some of the DX stations worked from the site include EI0CF, GI3PDN, PA0BWL, PA0LEG, GD0MRF/PA0SE, SM6PXJ, DK5PT and HB9DCE. This last contact was interesting. It would appear that the mass of high ground to the south of the site (up to 100m higher than the antenna and covering several kilometres) has very little attenuation at LF.

Although the LF station is not part of the exhibition, the museum is well worth a visit. It has all manner of working exhibits from our industrial past, including a narrow gauge railway with industrial steam, diesel locomotives and vintage buses, which you can ride on. Demonstrations of all types of skills and crafts are too numerous to mention here.

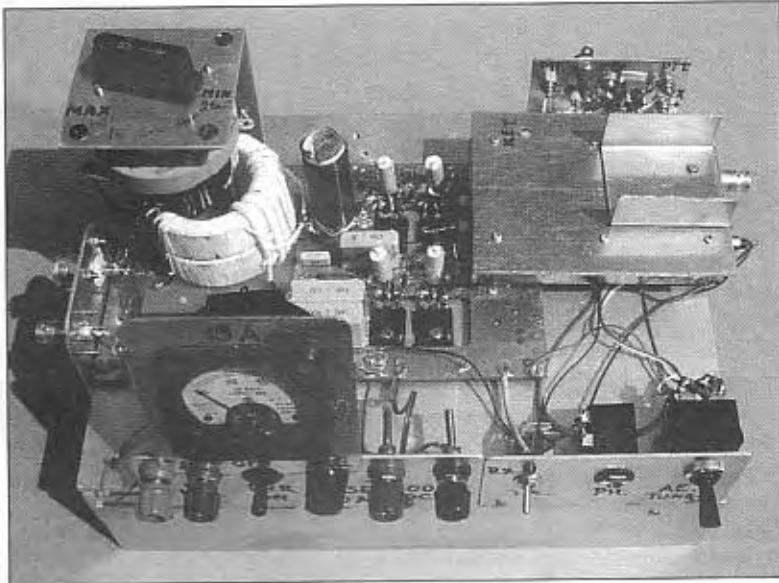
Fig 1.17: Operating position inside the vehicle. The variometer and matching transformer components shown strewn around the 'desk' on the right; part of the attempt to tune and match the antenna.



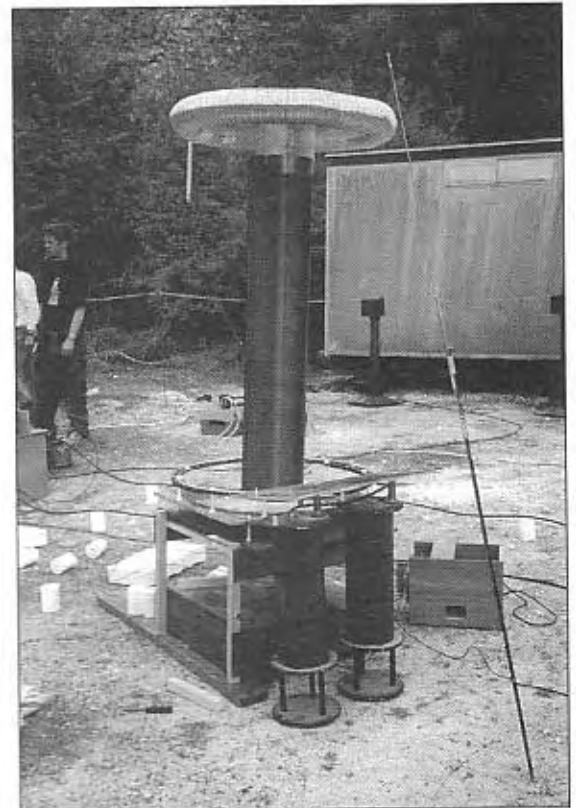
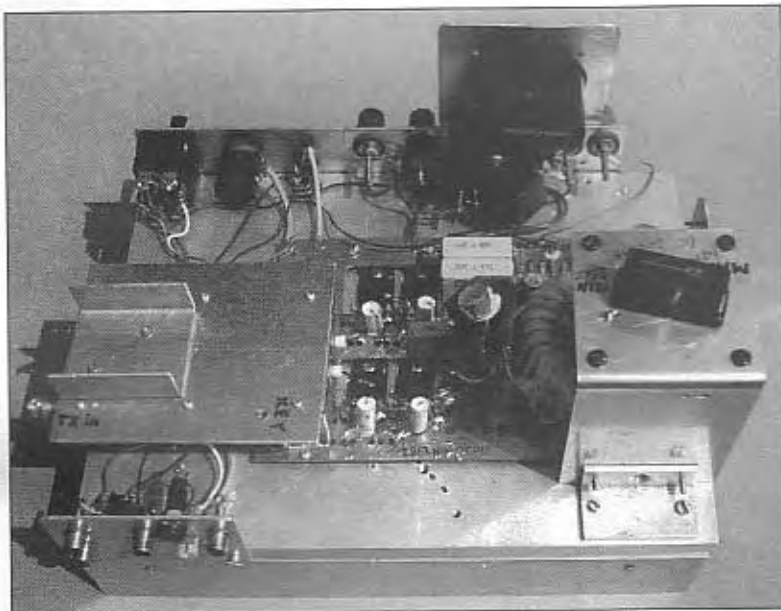
Photo Gallery

Included here, because of their general interest, are photographs that were not used in the text of this book.

LF transmitter built by Derek, G3GRO. It is based on a commercial BK Electronics audio amplifier.



(left): 1kW Class D transmitter built by Lech, G3KAU. This design uses the original GOMRF board with some modification.



A really big coil. A Tesla coil being set up at Amberley museum. The resonant frequency of this arrangement was around 170kHz.

(left): Rear view of the G3KAU kW transmitter.



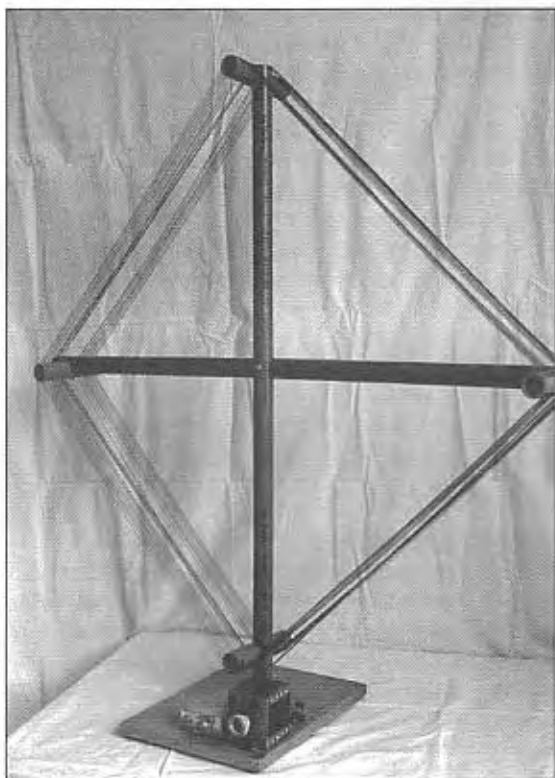
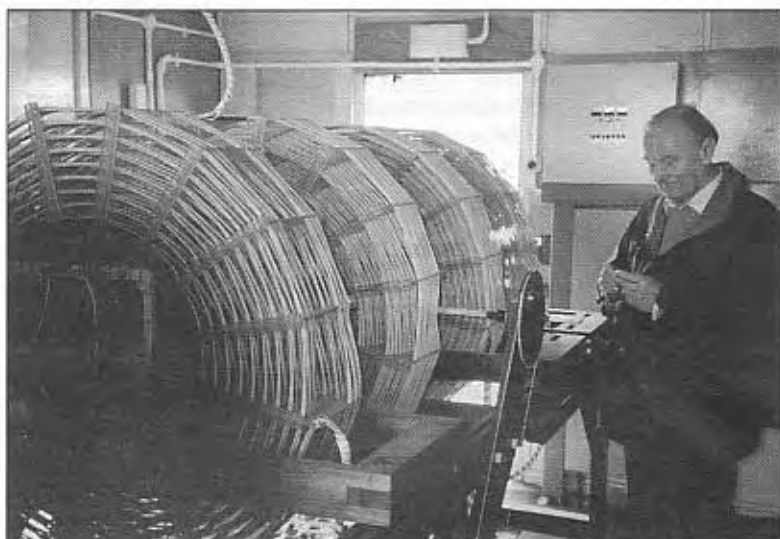
Gift-wrapped. The GW4ALG loading coil with weather protection.

The LF loading coils seen during the de-commissioning of the Decca station, near Lewis in Sussex. The coupling between the nearest coil and second coil along is automatically adjusted for changes in the antenna ground due to changes in weather. The nearest coil is mounted on a wooden frame that runs on glass ball bearings and the frame is moved via the pulley at the side of the coil assembly. Lech, G3KAU, contemplates the difficulties of fitting it all into his shack!

The G3LDO loading coil with weather protection. A plastic cover provides surprisingly good protection against rain, frost and snow.



Remote controlled variometer by GW4ALG. The remote control mechanism is driven by Lego (children's construction kit comprising plastic cog wheels etc). The complete assembly is built into a plastic storage container with a tight-fitting lid.



A well designed and constructed receiver loop antenna by PA0SE. The amplifier and tuning arrangement is fitted to the base.

2

Receivers

LF Receive Systems

The following is based on an article entitled *Optimising LF Receive Systems*, by John Moore, G4GVC. It has been edited to reflect two or three years' collective experience using LF and includes additional information on currently available receivers

The LF allocation provides us with a completely new challenge and the chance to experiment in a part of the radio spectrum previously unavailable to amateurs. Contrary to some popular opinion, these wavelengths are neither so large that it is impossible to generate useful signals from average suburban locations, nor are the frequencies so low that techniques are simple and fully understood, with no development work left to be done. Working on the new bands, one soon appreciates that these frequencies have some quite different properties to both audio and RF. New techniques still need to be developed and many old ones adapted for amateurs with limited resources to achieve optimum results in this part of the spectrum.

Good results can be obtained on receive with basic equipment such as war-surplus receivers (many of which cover this part of the spectrum), simple home-brew receivers, HF equipment with standard VLF converters and simple antennas. For those interested in pulling out weak DX signals from the very high LF noise, it becomes necessary to carefully optimise the receiving system in order to detect and copy these signals.

The following notes are a summary of personal, empirical observations over several years experimenting with receiving systems at LF. They are not intended to discourage those who want to experiment with simple receive systems, which can produce perfectly acceptable results, but are intended to encourage those interested in fully exploring these frequencies to develop more effective receive set-ups.

LF Weak-Signal Receive System Requirements

Optimum Sensitivity

The complete receive system; *ie* the receiver and antenna must provide enough sensitivity to receive the very weak amateur signals on these bands. Unless the station is reasonably local, say less than 50km away, the signals will be much weaker than the other signals audible on the band such as navigation beacons, data transmissions and time signals. Many of these are very strong indeed, so the ability to receive non-amateur signals well does *not* in

itself indicate adequate sensitivity. Even during the daytime, on clear frequencies, loud static crashes should be audible which affect the receiver AGC and frequently cause the S-meter to 'kick' – similar to listening to 160m on a humid summer evening. If these crashes are not plainly audible, the receive system is not sensitive enough and steps should be taken to improve the antenna and/or overall receiver gain. An optimum sensitivity must be achieved, however, as too much front-end gain can be disastrous, causing very strong non-amateur signals to overload the receiver with resultant cross modulation/inter-modulation problems, as described below.

Maximum Signal to Noise Ratio

At these frequencies, electrical noise sources, both in the shack and close to the antenna, can totally swamp a wanted signal. Typical noise sources are computers, televisions, switching PSUs, fluorescent and 'long-life' lighting, dimmer switches, multiplexed displays (eg the one in the rig!), alarms etc – all the things you are likely to have around the home. The obvious precautions to be taken here are:

- Minimising the number of possible interference sources left running whilst listening.
- Siting the antenna as far away from noise sources as possible.
- Making careful use of a directional antenna such as a loop to null out noise and interference, using screened or balanced circuitry where appropriate.

Too high receiver sensitivity, poor receiver gain distribution and/or receiver non-linearity can have disastrous results at these frequencies, where very strong adjacent signals and nearby broadcast stations are almost certain to cause overload, cross-modulation and intermodulation products in a poorly designed receiver (remember that the Long Wave starts less than 15kHz above the 136kHz band!) These problems cause effects such as high general background noise, broadcast modulation audible on other signals, 'phantom' data signals and signals 'keyed' by time pulses – all very common problems in poor LF receivers.

Except when using a very narrow-band, low-gain loop antenna, it is best to avoid front-end preamplifiers. Any noise generated by the preamplifier itself will be amplified by the rest of the receive system, but far more importantly, the preamplifier or following stages will almost certainly be overloaded by strong unwanted signals received on a good external antenna, even when filtering

is used. The best strategy, whether using a converter or general coverage receiver, is to reduce or even omit any gain until after the first mixer and concentrate on good hi-Q tuned circuits directly following the antenna to give maximum attenuation of out-of-band signals. Any lack in overall gain can then be made up in later stages where these problems are less troublesome.

Choice of the most suitable IF and AF bandwidths is very important. Most amateur signals on these bands are either CW or narrow-band data, so the use of good CW filters is more or less mandatory (a 'standard-width' SSB filter is wider than the whole 136kHz band and almost as wide as the 73kHz band!).

The IF filter bandwidth normally needs to be 500Hz or less, and something around 250Hz is preferable, provided it does not ring. Ringing is particularly troublesome on these frequencies owing to the high impulse static noise, and causes a further increase in noise and a reduction in signal readability. Receivers with the ability to cascade filters at different IFs and a 'slope-tune' control are particularly useful, as this allows the overall IF bandwidth to be reduced still further with a minimum of ringing. A very useful feature is to be able to swap carrier insertion oscillator to the other side of the filter (often called CW REVERSE mode). If a signal is suffering adjacent interference, swapping to REVERSE mode often shifts the interference outside the passband, or at least changes its pitch to something less intrusive. Analogue audio filtering may also be used with care although it often has a tendency to worsen readability when used excessively. Audio filters should not be used as a substitute for adequate IF filtering.

DSP audio filters can be effective on some signals but early units may not perform well in the presence of LF impulse noise (which may not conform to the filter's noise-elimination algorithm and may also upset the filter's operating threshold). No doubt later units are more effective in this respect.

High receiver local oscillator noise can cause problems, where there are often very strong in-band data signals present. The most obvious result of this is the inability to hear weak signals close to these strong signals, owing to oscillator noise sidebands. The use of well-designed crystal oscillators and DDS techniques are effective ways to minimise oscillator noise.

Antenna size, positioning, orientation and polarisation can all be used to maximise the wanted signal strength and minimise received noise and interference. The size of the antenna is always a compromise between achieving adequate sensitivity and minimising noise pickup and unwanted signal overload. Positioning the antenna as high and far away from noise sources as possible usually gives the best receive signal-to-noise ratio, but this may not be necessary with small frame-loop antennas, which can often give good results inside the shack. Loops in general have directional properties and should be carefully orientated in order to effectively null out noise and/or optimise the wanted signal. For the surface waves most often received on LF, vertical antennas are usually considered most effective, but these are also more prone to picking up electrical-noise. G4GVC found that horizontal wire antennas can work very well on receive at LF and may also provide enhanced reception of any distant sky-wave signals, particularly on 136kHz.

At LF, band noise varies considerably throughout the day, peaking during the evening and is at a minimum in

the early morning. Exact times change with the time of year, and these effects appear to be the result of both variations in propagation and the amount of noise-generating electrical equipment in use. Much of the operating is done in the mornings (often around 08.00 – 09.00UTC) when the band noise is low and signals are much easier to copy. There is some evidence of long distance skywave propagation in the evenings on 136kHz, and very distant signals may be audible then despite the higher noise levels.

Accurate Calibration

When searching for very weak signals, it becomes very important to have confidence that the receiver is tuned exactly to the desired frequency. For aural CW, the frequency calibration of modern receivers and transceivers with digital frequency readout is adequate. The normal procedure is to search using a fairly wide filter, say 500Hz, then to reduce the bandwidth to suit the receive conditions.

Some confusion may occur on receivers where the calibration does not take into account the position of the carrier insertion oscillator relative to the IF filter passband in the CW mode.

This can be resolved by checking the calibration of the receiver against one of the LF Standard Time Signals (MSF, HBG, DCF77 *etc*). With a transceiver this can be done by tuning on to the frequency standard, while at the same time keying the transmitter very slowly (it should not transmit outside the HF amateur bands). The transceiver should be on tune when the signal's beat note is exactly the same as the sidetone note.

CAUTION. If your transceiver is set for 'general coverage' transmit, or if you are using a transverter or converter with an IF in the amateur bands, you may transmit with damaging or at least unpredictable results when you key the transceiver sidetone. Make sure that you do not transmit when you use the sidetone to check receiver calibration.

Ideally the tuning accuracy should be better than 10Hz and, if possible, nearer 1Hz, especially when narrowband modes such as PSK or very slow CW are being used. This is now relatively easy to achieve at these frequencies with modern techniques

Future ultra-narrow-band techniques may require some sort of high stability, high accuracy master oscillator in the receiver (a TCXO, off-air standard, atomic standard *etc*).

Receivers

G4GVC settled on the Kenwood TS-850S as his main LF rig after having tried a number of different receivers over the years. Despite the receiver being specified only down to 100kHz, it appears to work well below 70kHz with excellent performance. John's enthusiasm for this receiver influenced other members of the LF group (including G3LDO) to acquire the TS-850 and no one was disappointed.

The following are comments from other LF operators users regarding receivers:

Rik, ON7YD, uses a Kenwood TS440 on LF, sensitivity is good and if the internal 20dB attenuator is used the IM behaviour is very good (using my 'big' antenna for reception).

Toni, HB2ASB, uses a Harris RF590 for LF reception, followed by a Timewave DSP-599zx. The Harris tunes

from 10kHz up to 30MHz and is available on the surplus market. It was in widespread use in many US civil and military services and was made in the USA from 1982-1998.

The advantages are:

- 1Hz read out
 - Very good stability with the built in oven (still within 1Hz after 2 years of service, without adjustment).
 - Slow tuning rate of 100Hz for 1 turn of the VFO knob.
 - Synthesized BFO tuning in 10Hz steps.
 - Adjustable line out for the PC sound card.
 - The receiver is fitted with the automatic preselector option and has an outstanding IMD and cross modulation immunity.
 - There are different AGC settings, including OFF.
- The disadvantages, and some remedies are as follows:
- IF blow by, owing to insufficient blocking of the filter control circuit (can be modified).
 - Too large wide a CW filter. Originally it was equipped with a 300Hz filter – but at -3dB this corresponds to a 500Hz -6dB filter as it is used in amateur transceivers. This filter has been replaced with a Icom FL 53A (250Hz at -6dB).
 - An uncomfortable VFO knob – replaced by one from a WJ8888.

The Timewave determines the audio bandwidth and this is set at 35Hz most of time. The built in noise reduction algorithm works quite well and attenuates the worst static noise (no NB option in the Harris).

An Icom IC765 is used as a back up receiver, which has the same sensitivity as the Harris. The Icom has a very good noise background but it is useless without the narrow CW filters (250 Hz in both IFs). This receiver is also used with the Timewave DSP.

Marco, IK1ODO, uses a Racal 1792, which has the following advantages:

- 1Hz tuning step.
- High stability frequency reference, and all the synthesizers derived from the same 1MHz signal.
- Synthesized BFO.
- Good AGC, can be switched off.
- Good dynamic range (about 102 dB 3rd order IMD at 14MHz, never measured at 137kHz) – ample choice of filters.
- IF pass band tuning.
- In addition it has no bells-and-whistles and is easy to repair and/or modify.

Laurie's, G3AQC, main receiver is an old Icom IC781, which has two excellent 250Hz CW filters. Gain is more than adequate on 137 and the spectrum display is useful for keeping an eye on band noise but too wide for anything else. The 250Hz filters are very good but sometimes too wide so an NIR-12 DSP filter is used, which can go down to about 60Hz.

The Icom IC756 PRO has very excellent digital filtering and the 50Hz setting is quite useable under all conditions, even with high QRN. The display of filter width and passband tuning is very useful but again the spectrum display is not too useful although it is somewhat narrower than the 781. The big problem with this receiver is that is a bit deaf on 137 about 18dB down on the old 781.

A Racal 1792 is used for band monitoring and for receiving QRSS. This an excellent receiver with 1Hz tuning steps.

A four-section bandpass filter (PA0SE design) modified for 137kHz is also used. This filter is followed by about 12dB of gain since the loop and EWE type ants have very low outputs. An emitter follower output stage is used so that more than one receiver can be connected to a single antenna.

Several LF operators, including **Vaino, OH2LX**, and **Alan, G3NYK**, are using the AOR7030 with good results. Coverage is from Zero to 33MHz, with 10Hz resolution. Optional 500Hz and 300Hz CW filters are available.

Rich, OM2TW, uses an Icom IC-761 as receiver on 136kHz and finds its performance excellent. It is fitted with four optional CW filters, 500Hz and 250Hz in 9MHz IF and 500Hz and 250Hz in 455kHz IF. For comparison an EKD300 German receiver was borrowed from OM3UU for a few months. It has a frequency range of 14kHz to 30MHz. This is a very good receiver with an excellent range of filters, including 700Hz, 500Hz, 250Hz and 50Hz.

Mike, G3XDV, uses an IC-706 Mk1. Although it will receive directly on 136kHz the sensitivity is poor, and would need a pre-amp to be useful. The receiver will operate on 73kHz, though the specification states the lowest frequency is 100kHz. In practice the IC-706 is used as a tuneable IF on 28MHz and used with the Datong converter. The frequency tuning steps and readout resolution is 1Hz – very useful on LF. The dual VFO allows instant very accurate frequency calibration by having one VFO on 60kHz and switching between the two. The 500Hz optional filter is adequate and there is room for a 250Hz filter. A socket provides a fixed audio level, which is ideal for Spectrogram or Spectran.

Derek, G3GRO uses the FT-990 transceiver as an LF receiver and finds that it is excellent. Ergonomically it is very good, with an uncluttered layout of controls and no unnecessary 'bells or whistles'. Options include very good filters – 250Hz, 500Hz, 2kHz and 2.4kHz plus AM and NBFM filters. The audio DSP works well. Signal handling is also very good. The sensitivity remains good right down to 136kHz which is not the case with some receivers.

David, G0MRF commented that he felt that the FT-990 performed better during the Puckeridge experiment (see previous chapter) than his fairly new Icom IC-756 both of which were being operated with antenna inputs in parallel. Like the IC-756, the FT-990 has twin antenna inputs to the receiver, which is useful. A couple of minor criticisms: The AGC discharge time is too long even on the 'FAST AGC' position and it takes some time for it to recover after a crash of static. Often it is better to run without AGC but the QRN can be painful on the ears. However, PA0SE's diode limiter, described at the end of this chapter, can remedy this. In comparison with say a TS-940; the noise limiter is pretty ineffectual.

Lech, G3KAU also has a FT-990 and he shares these views.

Practical LF Receiving Antennas Frame-Loops

The frame-loop is often the first type of receive antenna

tried on LF. It is small, fairly simple to construct, can be used in the shack, outside (if suitably weatherproofed) and when 'portable', has useful directional properties and can give good results if designed and operated carefully.

Details of the construction of a loop for 136kHz are given later in this chapter.

Outdoor Wire Antennas

There is a consensus of opinion that large outdoor receive antennas on LF are unnecessary as reception is ultimately limited by band noise, not signal strength. Whilst this may be the case for reasonably strong signals, some amateur signals are so weak that a good antenna is required to receive them at all.

G4GVC feels that a large, quiet, outdoor antenna is required when looking for the weakest signals. His wire antenna, approximately 60 metres long and 8 metres high, is electrically tiny at these frequencies; however the dimensions are not critical. The aim should be to get as much wire in the air, in the clear and as far away from noise sources as possible. Obviously this will be limited by what space is available at each individual location.

Another theory suggests that vertical antennas work best for the surface waves used at LF. In practice it has been found that a mostly horizontal antenna works superbly for receiving weak amateur signals on these bands at ranges varying from tens of kilometres to over 1700km, whether transmitted from verticals, loops or even ground electrodes!

G4GVC has found that it is very important to resonate the antenna with a high- Q series-tuned circuit as shown in Fig 2.1 (although matching the antenna to the receiver

input is not critical). This not only provides a very large increase in wanted-signal strength but also strongly attenuates the out-of-band signals which are likely to cause receiver overload problems with a good outdoor antenna.

A 7mH inductance will probably be a good starting point for 136kHz. It is well worth experimenting with different types of inductor if you have a junk-box, because of large variations in received signal strength when using different inductors in this circuit.

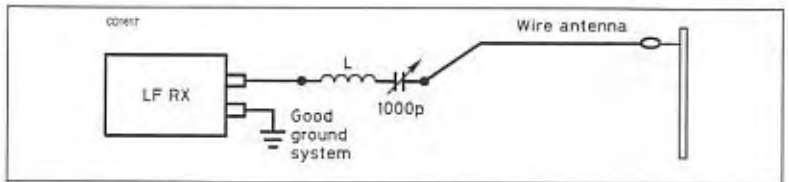
The importance of a good ground system when using an antenna of this type on these frequencies cannot be over emphasised. Your ability to receive weak signals will be limited by the effectiveness of your ground, so put plenty of effort into improving that, especially if you live in an area of poor ground conductivity.

Other Receive Antennas

The recent decommissioning of the Decca navigation system transmitters have given UK amateurs some experience of using very large 100m high antennas for transmit and receive.

Whilst these antennas worked fine on transmit they often gave problems on receive owing to overloading of the receiver front end by the strengths of the signals. The use of selective filtering at 137kHz and appropriate attenuation resolved the problem in at least one case.

Fig 2.1: Example of a simple circuit for resonating an antenna to improve reception at LF



LF Converter from 136kHz to 14.136MHz

By Hans-Joachim Brandt DJ1ZB

This converter was first operated without any pre-amplification, just employing a tuned circuit between antenna and mixer input. The choice of 14MHz as an IF was because it has the advantage that older receivers without WARC bands can be used as a tuneable IF. When this converter was connected ahead of a typical 14MHz transceiver the sensitivity was just sufficient, and the receiver noise peaked when the LF input circuit was tuned, but most LF amateur signals did not produce an S-meter indication. Because of this a pre-amplifier was added so that weak DX signals could be received during the low-noise winter months. The circuit is shown in Fig 2.2.

Mixer Circuit

Experience for this design was gained in the field of QRP construction on the HF bands. It employs the 74HCT4066B as a doubly balanced mixer, which is used mainly in direct conversion receivers. Because of the relatively high on-resistances of the internal FET switches, compared to typical diode mixers, it is recommended that it is used at a higher impedance level, such as 330 ohms. R13 may be regarded as the source resistor for the mixer, and the mixer output contains a symmetrical bridged T-network providing a broadband 360Ω load resistor and a

narrow output in the 14MHz range.

All trimmer capacitors have to be adjusted for optimum signal transfer at around 14.136MHz. Whilst L3 is an Amidon core, moulded chokes can be employed for the 22μH coils. This network was designed using the ARRL Radio Designer Programme.

The original output transformer core is a Siemens B62152-A0007-X001 (ferrite material 80K1) but may be replaced by an Amidon BN-61-2402 or equivalent. The turns ratio of (2x4):3 given in the circuit provides the match from 360 ohms symmetrical to 50 ohms asymmetrical.

One advantage of this balanced mixer output is the good cancellation of the 14MHz oscillator in the output.

Oscillator Circuit

The 14MHz crystal is excited in the series mode using two gates of a 74HCT00. The separate DC biasing combined with RF feedback around these gates will allow crystals in the range of 3 to 20MHz to be used without further changes. The other gates are used as buffer and inverter to generate the two oscillator signals needed with a phase difference of 180 degrees. The trimmer capacitor must be adjusted to exactly 14MHz with the help of a

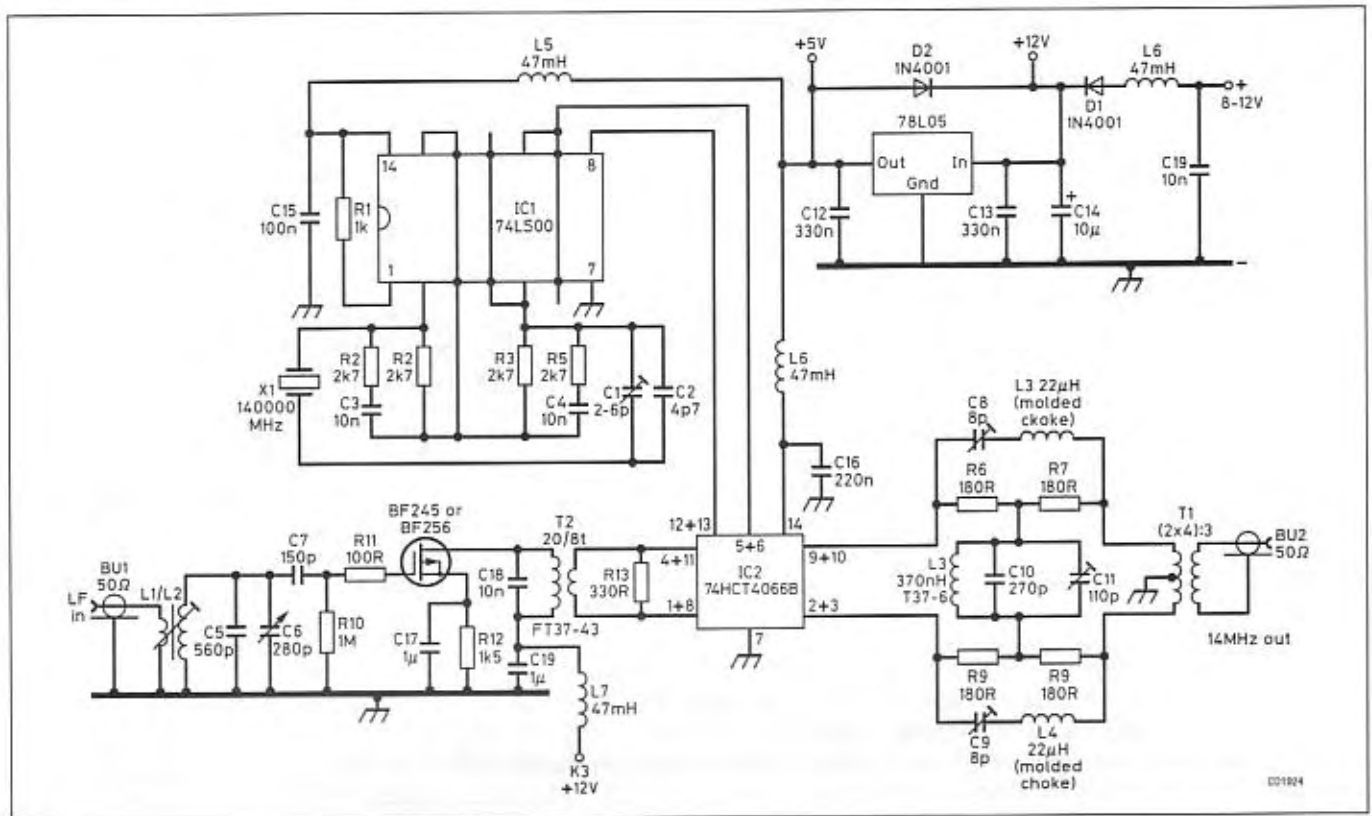


Fig 2.2: Circuit diagram of the DJ1ZB LF converter.

good counter. To make the best use of the good oscillator frequency cancellation of the mixer, the 14MHz oscillator is housed in a separate metal case screwed to the converter case. Three wires only are used leading in through a small hole; a +5 V supply and two slightly twisted wires for the push-pull oscillator outputs.

The Pre-amplifier

For the pre-amplifier an FET cascode circuit was tried first, but resulted in an unnecessarily high gain of over 30dB. A lower gain amplifier with an output resistance of just 1500-2000 ohms has been found sufficient. Therefore the design could be simplified considerably and now consists of a single gate FET (BF 256, BF 245 or similar) with a loaded drain resonant circuit with a Q of 12 to 18. The inductance of the drain coil was determined using an inductance meter, to get the broad resonance optimum at 136kHz while incorporating the tolerances of the Amidon FT37-43. The output resistance of the amplifier is controlled by the turns ratio (20:8 in this case) and the 330Ω input resistor of the mixer circuit. Employing a broad resonant circuit here has two advantages: fewer turns are needed compared with a broadband transformer, and the 10n capacitor is short circuiting the noise which may exist on the image frequency (at 28.1MHz) of the mixer.

The input circuit of the pre-amplifier first tuned using a Japanese foil variable capacitor with all sections paralleled, resulting in a maximum capacity of 280pF. The coil former is an old-fashioned slug-tuned version with four sections, which regrettably has almost disappeared from the market. This coil requires 4 x 80 turns of RF Litz wire for resonance (in conjunction with the 560pF fixed capacitor in the circuit). Just 7-8 turns on the 'coldest' turns section were sufficient for the coupling link to the LF transmitter antenna tuned to 50 ohms. At LF there may be a problem owing to coupling between these coils and 50Hz fields in the shack. For this reason the 150pF capacitor and the resistor from gate to ground are used to form

an RC high-pass to attenuate this 50Hz energy before it reaches the FET input. The 100Ω series resistor before the gate is a protective measure against VHF oscillations.

Intermodulation in LF Receiver Input Coils

It must be emphasised here that the coil type at the input of the LF receiving section is of utmost importance. There had been reports by ON7YD that (at least in Europe) the high levels of RF from nearby broadcast and commercial may saturate pot cores in such an input filter, causing a mess of intermodulation products in which the weak amateur signals will disappear. Simple air coils or slug-tuned coils like those employed in former radios do not tend to saturate so easy and are therefore preferred. If only pot cores are available for the design of LF receiver coils, the simplest way would be to remove one half of the core to create a sufficient 'air-gap' to avoid saturating. The data of such a modified pot core must be tried out by measurement.

Receiver Amplifier and Loop Antenna for 136kHz

By Andy Guye F6CNI

Many amateur radio transceivers cover the low frequency bands but often their sensitivity is very poor. The options are to make a converter or construct an amplifier to overcome the lack of sensitivity at the receiver. This amplifier design provides 40 to 45dB gain and is designed to work with a loop receiving antenna.

The Loop Antenna

The loop antenna comprises seven turns of 1.5mm² insulated electrical wire, see Fig 2.3. The loop is a square configuration, eight metres per side, mounted on a 12-metre high pole; the sides are supported by thin nylon rope. The total length of the wire in the loop is 224m, about 0.1 wavelength at 136kHz. The antenna is connected directly to the high impedance input of the amplifier to preserve the Q of the loop, which is approximately 500. The amplifier is located next to the loop and the power for the amplifier is fed down the 50Ω cable that connects the amplifier to the receiver.

The antenna has maximum sensitivity along the axis of the nylon rope supports, with two deep nulls at the sides. The loop will need to be orientated to the direction of most interest so it is a good idea to plan the location of the antenna.

The Amplifier Circuit

The amplifier is mounted in a closed aluminium box, which in turn is located in a plastic weatherproof box. This is mounted at the base of the loop so that the leads to the amplifier are kept short.

The loop antenna, is connected directly to the gate of TR1 via a 1500pF capacitor, see Fig 2.4. The loop is tuned with a 200pF air-spaced capacitor Ca (my loop resonates at 137kHz with 140pF). If variable gain is required then the MPF102 at TR1 could be changed for a 40673, or similar, with a potentiometer in gate 2 to obtain the variable gain.

The filter circuit, resonant at 136.8kHz, comprises three top-coupled parallel tuned circuits.

L1 and L3 are 614μH and comprise 23 turns of 0.3mm OD enamelled copper wire on LTT-FN107-1005 orange core. This inductor is tuned with a 2200pF capacitor.

L2 is 75μH and comprise 33 turns of 0.3mm OD enamelled copper wire on an AMIDON FT50-61 core. This inductor is tuned with an 18nF capacitor. The values of the parallel capacitors may have to be changed slightly to get resonance, which is set at 136.8kHz.

The loop is located about 15 metres from the transmitting antenna, which has an unpredictable influence on the loop. The problem was solved by using relays to discon-

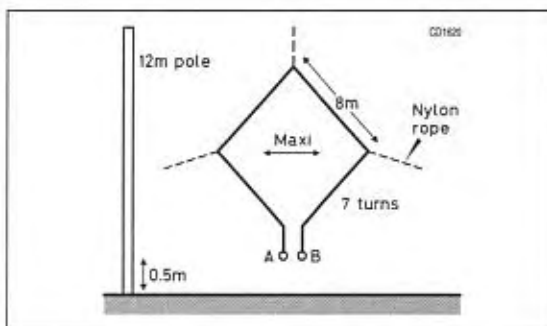


Fig 2.3: The loop antenna comprises seven turns of 1.5mm² insulated electrical wire.

nect the base of the transmit antenna from the 3.2mH loading coil on receive and to short out the receive loop connections to the receiver amplifier on transmit.

The coil relay is directly fed with the 12V coming from the coax during receive periods. During transmit periods there is no DC applied to the amplifier, so the amplifier is switched off, the relay de-energised, and the antenna input is connected to ground.

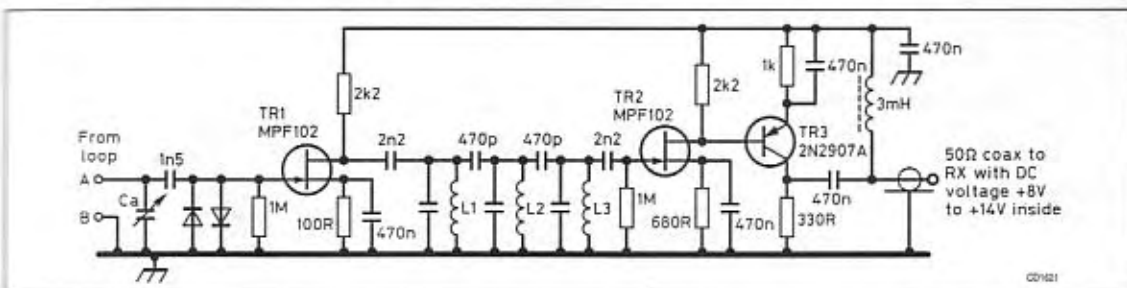


Fig 2.4: Loop amplifier circuit.

A 136kHz High Performance Receiver Converter

By John Moore G4GVC

No claim is made for the originality of the core circuit, which is taken from notes by G4COL in 'Technical Topics', *Radcom* April 1998. This converter is built exactly with the original values and on the 136kHz band it works exceptionally well. Those who have built this converter report that it is much better than the commercial Datong converter (which was not designed for this sort of service).

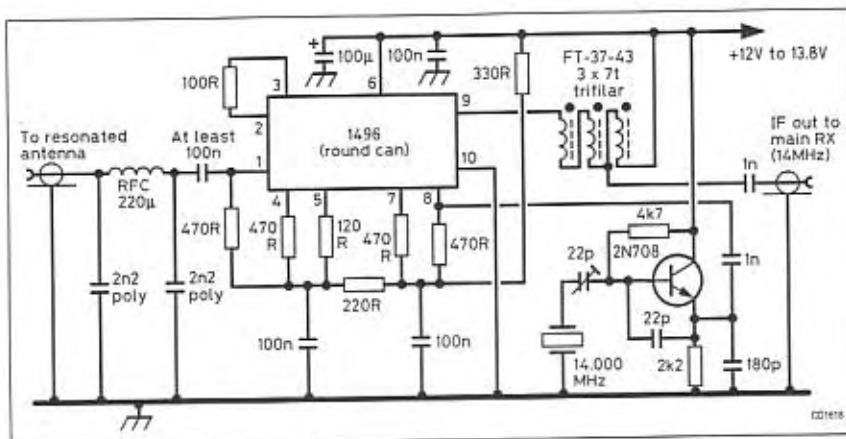
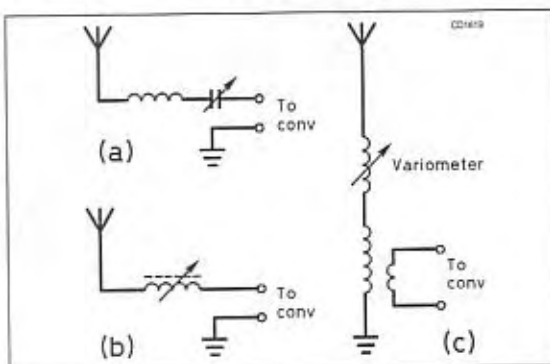


Fig 2.5: With a really good antenna tuner this design will produce excellent results from below 5kHz to the medium wave band.

The circuit

The circuit is shown in Fig 2.5. The pin numbers refer to

Fig 2.6: Antenna resonating circuits for the LF converter. Simple end-fed types are shown in (a) and (b). A higher performance resonating and input circuit is shown in (c).



the round can 1496 IC only. General purpose transistors will not provide sufficient oscillator injection.

A 10MHz output version can be made by simply substituting a 10MHz crystal and increasing the output transformer to 3 x 10 trifilar turns on the same FT-37-43 core.

The input pi-circuit is a very simple LPF, so to achieve the excellent intercept point potential an antenna resonating system must be used as shown in Fig 2.6. Anyone plugging a random wire straight into the converter will be disappointed.

All comments regarding optimizing antenna and ground systems in described in the beginning of this chapter apply here.

Noise Cancelling at LF

by Derek Atter G3GRO

The ability to resolve long-distance weak CW signals is often compromised by noise. This can be local electrical noise or sidebands of the wide-band pulse transmissions of the Loran-C navigation aid. These unwanted signals frequently determine the noise floor.

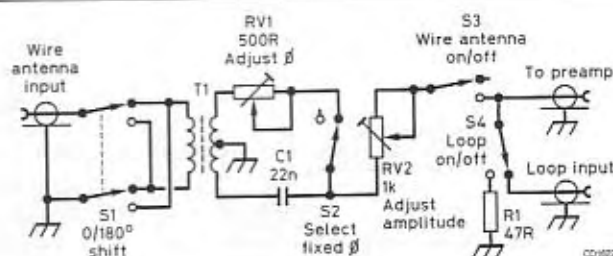
The first line of defence is to switch off the AGC, use as narrow a bandwidth as possible and employ DSP techniques. However, this is often not totally effective.

Several alternative methods can be used to combat such interference, including directional receiving loops, synchronous pulse blanking and noise cancellation.

The basic 136kHz canceller has only five passive components – see Fig 2.7. The signal from the omnidirectional wire antenna feeds an adjustable phase shifter with the output from the phase shifter being simply fed directly in parallel with the input from the loop antenna via an iso-

Fig 2.7. Basic canceller.

S1 is a two-pole c/o toggle; S2, S3 and S4 (not essential but add to operating convenience) are one-pole c/o. R1, 47k; C1, 22n; RV1, 500R (linear Cermet or carbon); RV2, 1k (linear Cermet or carbon); T1 is a trifilar wound toroid three times 18 turns of 28 or 32swg on a ferrite core (13.25mm diameter Philips C85 material or Fairite FT-50-43, or similar).



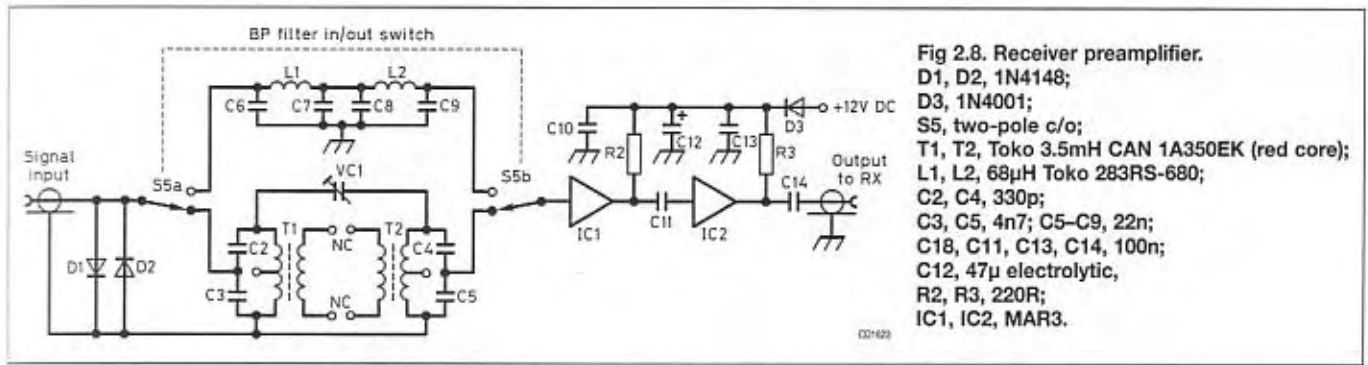


Fig 2.8. Receiver preamplifier.

D1, D2, 1N4148;
 D3, 1N4001;
 S5, two-pole c/o;
 T1, T2, Toko 3.5mH CAN 1A350EK (red core);
 L1, L2, 68 μ H Toko 283RS-680;
 C2, C4, 330p;
 C3, C5, 4n7; C5-C9, 22n;
 C18, C11, C13, C14, 100n;
 C12, 47 μ electrolytic,
 R2, R3, 220R;
 IC1, IC2, MAR3.

lating variable resistor RV2 which controls the amount of omni-derived signal being combined with that from the loop. A changeover switch S1 at the input to the phase shifter allows a 180° phase shift to be added to that of the phase shifter. Capacitor C1 and variable resistor RV1 together with the trifilar-wound transformer form a variable phase shifter with a range of approximately 180° and reasonably constant amplitude over its control range. The variable phase control RV1 does not quite reach 0° at one end of its travel and so switch S2 is provided which allows a fixed selection of the 0° position.

This simple signal combiner arrangement is based on the premise that the signal from the large omnidirectional antenna will normally be much larger than that from the loop and thus allows the omni signal to be fed simply in parallel with the loop output from the phase shifter via a reasonably high value of resistance in RV2 without adversely affecting the loop signal level. Adjusting the RV2 in conjunction with phase control RV1 allows the unwanted signal to be nulled out. It is also useful to be able to switch off the omni or loop inputs individually via S3-S4.

Most general-coverage receivers tend to be somewhat short of gain at LF and although the signal from the main LF antenna is usually adequate, a loop antenna – even a relatively large one – may need some additional gain in order to provide sufficient signal level from the interference source to allow cancellation to be effective. A suitable preamplifier, Fig 2.8, has two stages of amplification and is preceded by a band-pass filter.

The bandwidth of the input band-pass filter is about 3.5kHz. Switch S5a/S5b is an optional facility that allows this filter to be replaced by a low-pass filter with a 3dB cut-off frequency of around 180kHz so that the preamp can be used from a few kilohertz up to above 200kHz for general LF use. If the preamp is to be used only for 136kHz then components S5a, S5b, C7, C8, C9, C10, L1 and L2 can be omitted. Two MAR3 Minicircuits ModAmps are used in series, each having a gain of 12dB and a -1dB saturation point of +10dBm together with a good noise figure. They also have a nominal input impedance of 50 Ω .

The choice of two MAR3 gain-blocks, rather than a single MAR6 with a similar total gain, is dictated by the better signal-handling capability of the MAR3. The -3dB bandwidth of the input band-pass filter can be adjusted to approximately 3.5kHz by means of the small top-coupling trimmer capacitor VC1 between coil L1 and L2 to give a flat top response or slightly over-coupled double-hump response, after first peaking the cores of L1/L2 with loose coupling. Diodes D1/D2 at the input protect against unwanted transmitter power.

A strong carrier can be nulled by typically 30dB but the

phase and amplitude controls then become quite critical to achieve that depth of null. If the aim is to use the canceller on very strong carriers *etc.*, it may be beneficial to split the amplitude adjustment RV2 into separate coarse and fine controls. Avoid wire-wound potentiometers.

The omnidirectional input to the canceller system is normally the transmitting antenna fed to the canceller system via the transmit-receive switching relay.

A 6m (20ft) square loop antenna is used as the 'noise antenna'. It uses 1.5mm stranded insulated wire to keep the RF losses low. The individual turns of the loop are just bunched together and suspended at each corner via an insulator. The loop is resonated by two capacitors in series across the loop, mounted at the centre of the lower leg providing capacitive tap matching to the output coaxial cable. The 'earthy end' capacitor is 47nF in parallel with 10nF across the 50 Ω feeder; the 'top' capacitor has a value of 7.65n made up of several individual capacitors in parallel, one of which is a 500p compression trimmer. These are mounted in one of the small sealed kitchen food storage containers. The loop may be tuned to resonate at 137kHz by injecting a signal generator into the 50 Ω output port, connecting an oscilloscope or RF millivoltmeter across the loop via a high-impedance probe and adjusting the capacitance for resonance.

The loop interface box may, if desired, include a small 12V relay and RF choke (not shown in the diagrams) to allow DC to be fed down the coaxial cable to the loop so that a second loop running at right-angles can be remotely selected from the shack.

The GW4ALG Noise Canceller

By Steve Rawlings GW4ALG

Using small LF antennas from an urban or sub-urban location is a challenge in itself. Just one local source of noise can make it impossible to hear any amateur signals on the band. It was local noise that prompted me to build my first noise canceller.

The principle is quite easy to understand: Obtain a sample of the interference signal; amplify it; and then couple the 'noise' signal into the receiver so that it exactly cancels the noise signal picked up by the main antenna.

Initial Results

Initial results with a noise canceller on local sources of interference were most encouraging.

The circuit in Fig 2.9 is based on an idea by VK5BR (*RadCom*, March 1993, p34). The RF loss through the

canceller (between the Main Antenna port and the Receiver port) was found to be about 8dB at 136kHz. Under test conditions, a sine wave can be nulled by greater than 50dB. My strongest noise source (possibly a light dimmer), which, since mid-1998, has prevented me operating in the evenings can now be reduced from S9 to less than S5. And the number of significant noise sources at my QTH has increased from 1 to 3 over the past 4 months

An untuned length of wire, about 10 metres long, located in the loft [attic] in proximity to mains wiring seems to be make a good noise sense antenna for nulling out interference received on my vertical antenna. This 10m wire does not, however, always provide sufficient noise to null out interference heard while using my 60m single-turn

Fig 2.9: Circuit diagram of basic noise canceller.

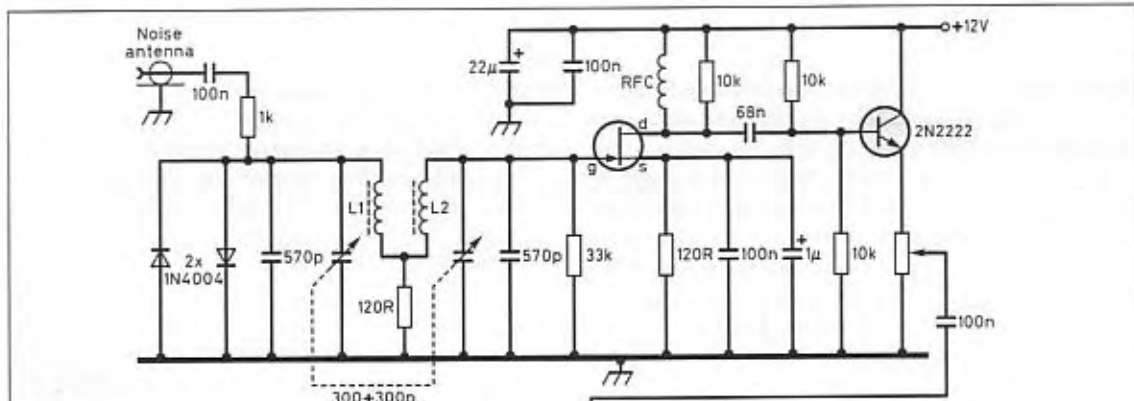
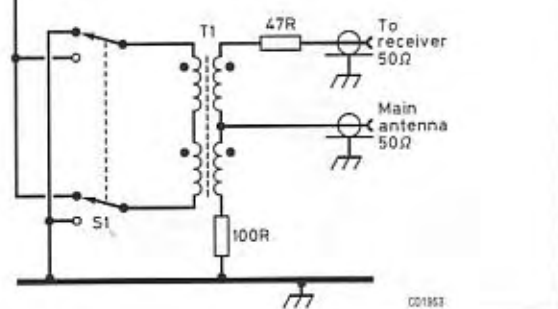
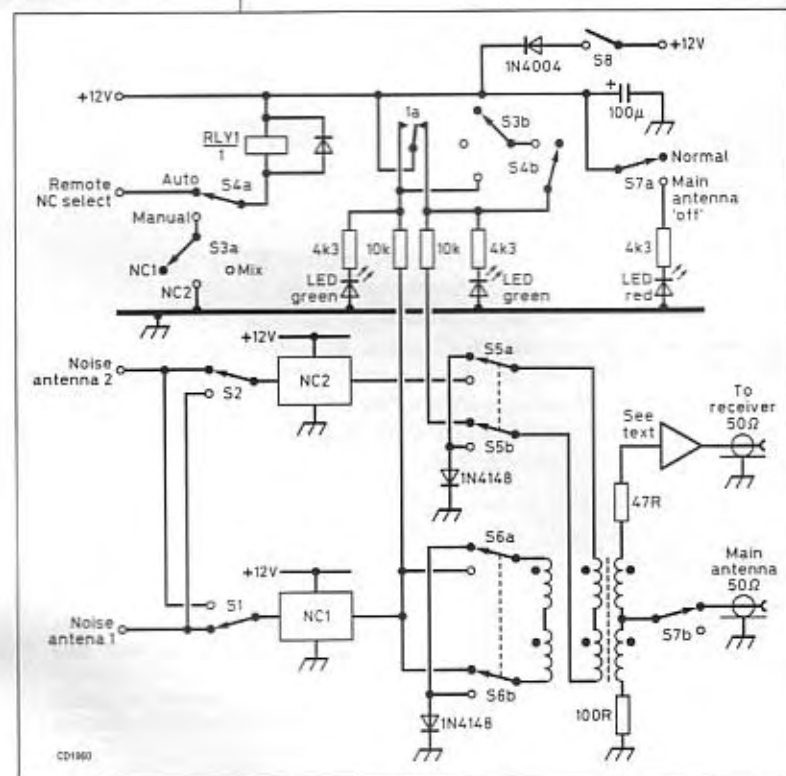


Fig 2.10: Circuit diagram of switched noise cancellers.



delta loop antenna. Only a combination of phase/amplitude settings, which take about 30 seconds to optimise, can null out one interfering signal at a time, for a given receive antenna.

Basic Noise Canceller

The original version of my VK5BR noise canceller, adapted for use on 136kHz, is shown in Fig 2.9. L1, L2 and T1 were made using 25mm OD 3C85 ring cores and wound with wire obtained from stripped-down internal telephone cable. The wire is plastic coated with a conductor size of 0.020 inch (25SWG, or about 24AWG). L1 & L2 are wound on separate ring cores and each is wound with 28 turns (see 'Alignment Procedure' below). T1 uses 13 (±2) turns, quadfilial wound at about one twist every 20mm. The variable capacitor provides a variable phase shift of 0-180 degrees, and SW1 provides a further 180 degree phase shift, permitting an overall phase change of

0-360 degrees. The potentiometer is used to adjust the amplitude of the noise signal. In practice optimum settings take several seconds – but the results are well worth the effort!

Alignment Procedure

The values for L1 and L2, together with the 570p of fixed tuning capacitance, will provide the required phase shift of 0 to 180 degrees. (If using a 200 + 200p variable capacitor, L1 and L2 should be wound with 34 turns, and the two 570p capacitors should each be changed to 400p.) But variation between different cores will mean that some adjustment will be required to ensure that the required phase shift range will be obtained.

Firstly, disconnect the variable capacitor wire to L2, connect the noise antenna and monitor the noise output on the station receiver. Noise on 137kHz should peak with the variable capacitor (now tuning L1 only) at about the centre position. (Note that the 'Q' of the circuit is quite low, so the tuning will be very 'flat', with a broadband frequency response.) If the noise gets stronger at maximum capacitance, increase the value of fixed capacitance. If the noise gets stronger at minimum capacitance, decrease the number of turns on L1, one turn at a time, until resonance occurs near the centre position of the variable capacitor. You may need to remove up to 6 turns from L1. When this has been done, re-connect the variable capacitor wire to L2; and then disconnect the variable capacitor wire from L1. Align as before.

Using the Noise Canceller

The useable bandwidth at the optimum cancellation settings seems to be about 600Hz, depending upon the strength (level of annoyance) of the QRM. The optimum apparent bandwidth cancellation decreases for an increase in the level of interfering signal.

Initial tests resulted in QSOs with HB9ASB, GD3XTZ/P, OH1TN, PA0BWL, GD0MRF, GB2CPM, G13PDN, ON7YD and SM6PXJ to name but a few. None of these evening QSOs would have been possible without the use of the noise canceller – the noise level was in the range S8 – S9 throughout.

A Pair of Switched Noise Cancellers

Because the basic noise canceller requires different settings for each of my two receive antennas (loop antenna/experimental vertical), I found myself having to adjust the noise canceller quite frequently. The current arrangement involves two switched noise cancellers in one box. A sense signal from the antenna switch to is used to auto-

matically select the appropriate noise canceller for the chosen antenna. Also, I have now introduced a 'mix' facility so that it is possible to cancel two different QRM sources simultaneously. The circuit diagram is shown in Fig 2.10.

Note that an extra winding has been added to T1 to couple the output of the second canceller. To make it easier to wind the transformer, I used a 42mm OD 3C85 ring core (a 25mm core would probably have worked just as well). To prevent interaction between the phase and amplitude settings of NC1 and NC2, avoid selecting the same noise antenna for both noise cancellers. Also, during construction, use 'point-to-point' wiring around SW1 and SW2 to reduce capacitance between the two circuits. In the 'Auto' mode, noise canceller 'NC2' is automatically selected when the main station antenna switch applies an earth (ground) to the terminal 'Remote NC Select'.

Of course, one thing leads to another! The 8dB loss between the Main Antenna port and the Receive port of the hybrid transformer meant that the usual S9 signals were only just reading S8 on the bargraph-type S meter of the FT707. To compensate for this loss, a receive pre-selector was constructed and connected at the Receive port as shown in Fig 2.10. A resistive pad was used at the output of the pre-selector to set the overall loss through the noise canceller to 0 dB.

Views of the front and rear of the noise canceller are shown in Fig 2.11 and Fig 2.12 respectively.



Fig 2.11: Front view of the GW4ALG noise canceller, showing the controls.

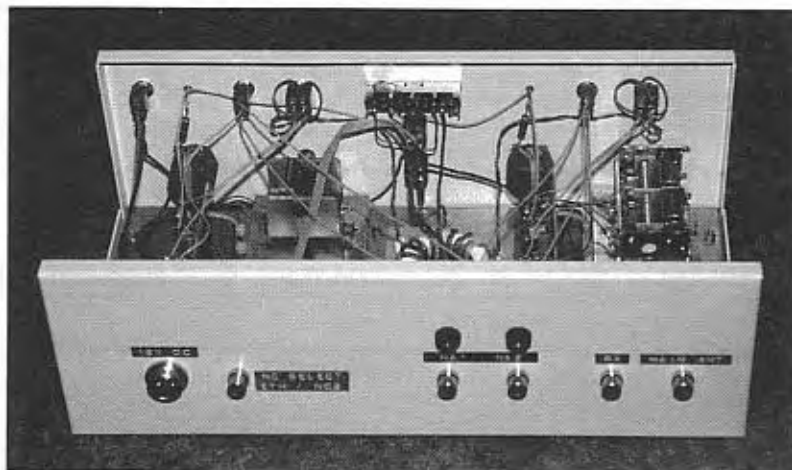


Fig 2.12: Rear view of the GW4ALG noise canceller, showing the connections.

An Audio Clipper for Headphones

by Dick Rollema PA0SE

On LF I use a simple audio signal clipper to bring all signals in the headphones to the same level. It may not improve readability but makes listening much less tiring by limiting all sorts of crashes and other impulsive noises. The clipper is now always switched in and always in use.

The clipper consists of two germanium diodes type OA85 connected anti-parallel over the output of the receiver. To be effective, both the voltage and the impedance at the point where the diodes are connected must be sufficiently high. I use passive audio bandpass filters between receiver and headphones, and the diodes are connected across the first tuned circuit of the filters where impedance and voltage are high enough (do not put the clipper behind the filter.)

The system will not work in a low impedance circuit with low resistance headphones. One solution may be to find a suitable point for the diodes in the audio section of the receiver; for example the voltage across the volume

control may be high enough for the purpose.

Another solution is to use a pair of old valve-type audio output transformers (for example with a ratio of 5000 to 8 ohms or thereabouts) connected back-to-back with the high impedance windings connected together with the diodes connected at that point. One of the low impedance windings is connected to the receiver output and the other to the headphones.

I also tried anti-parallel diodes in series with the receiver output, as advocated by Jan Smeets, ON4ASZ/EA3DPD ('Technical Topics', *RadCom*, March 1998) hoping that they would improve readability of weak signals in noise. The idea was that the output level of the receiver would be adjusted so that signal+noise would just overcome the threshold voltage of the diodes and the noise alone not. But it did not work. In retrospect, this was understandable, considering the wide amplitude variations of noise.

Outline Design for a Narrow Bandwidth LF Receiver

By Andy Talbot G4JNT

This receiver is proposed as an outline concept for very narrow bandwidth reception at LF frequencies, and whilst it is not presented as a fully working tested project, several of the critical individual sections have been built and tested and should fit together as described. The concept grew out of the breadboarding of a ladder filter made from 32.768kHz watch crystals that showed that a 1Hz bandwidth was quite feasible.

The aim is to provide a tuneable LF receiver with filtering to a narrow bandwidth to eliminate very close adjacent interference, such as Loran spectral lines, with a sampled digitised output at a very low data rate consistent with filtering to this bandwidth. A rate of around 10 samples per second proves more than adequate for a signal of this bandwidth, allowing the aliased products outside a 5Hz bandwidth to be considered insignificant. By generating the digital data at such a low rate, the use of extra DSP or soundcard hardware is not needed, the data being fed directly to a PC via the serial port. This releases a lot of the PC's potential processing capability for actually processing the signal and displaying results, rather than mixing down or decimating an audio tone; it also allows very slow and old machines to be used.

An essential requirement for the receiver was for *all* internal frequencies to be derived from one 5MHz master oscillator, which can then be generated by an external high stability source.

The complete proposed design is shown in Fig 2.13.

The front end is conventional, but note that the image response of 137kHz, for an LO tuning above the signal

frequency, falls very close to the Droitwich BBC transmission at 198kHz and so will need considerable attenuation; of the order of 70dB for many UK operators. Low side LO tuning means that the image response falls in the band 70 - 72kHz, needing a high pass or bandpass filter. A bandpass filter covering 120 - 140kHz is shown. Filtering for other RF bands needs to be chosen appropriately.

A MAR-2 modamp provides low noise amplification before the mixer, which is an SRA-8 packaged device having a lower frequency limit of 10kHz. The local oscillator consists of a Direct Digital Synthesiser module, tuning 32,765.5Hz above or below the wanted signal. The DDS module is not covered here as it is a standard application circuit for devices such as the AD9850 and AD9832 already in use elsewhere on the LF bands. By clocking the DDS input directly at 5MHz instead of the more usual many tens of MHz, a tuning resolution of 1.16MHz is achieved, up to a maximum output frequency of around 2MHz. Other suitable mixer types may be substituted here, but dynamic range of the LF spectrum in the wanted region needs to be considered in selecting a suitable device. An L match network converts the mixer output impedance of 50 ohms to the filter terminating value of 56k Ω .

The selectivity is provided by a prototype ladder filter made from watch crystals with a parallel resonant frequency of 32.768kHz. The centre frequency of a ladder filter is set by the series resonant frequency of the crystals, and this is lower than the parallel resonant value. A

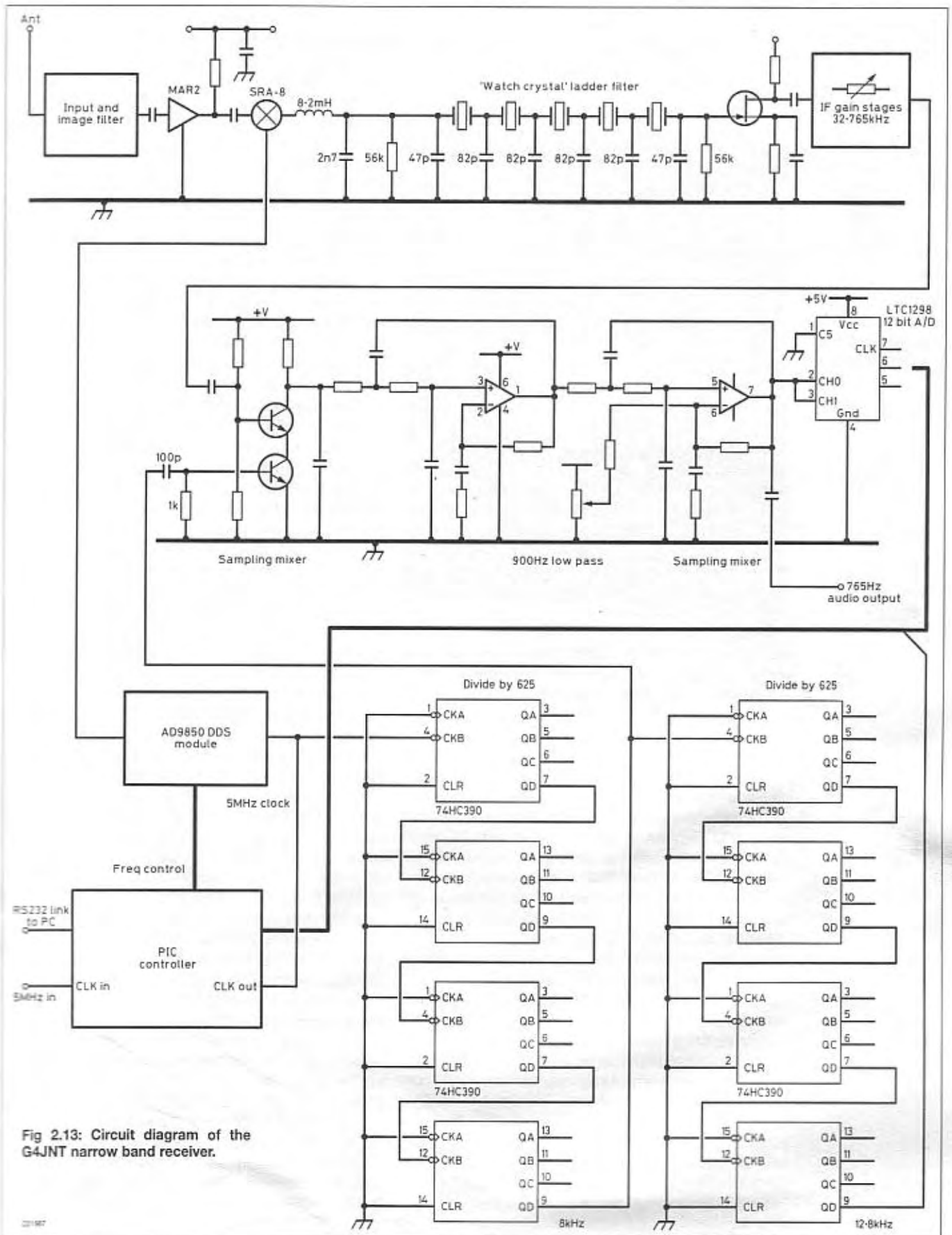


figure of 32765.5Hz was measured and this value then dictated the frequency for the IF. Watch crystals have a much higher internal resistance than higher frequency crystals in the MHz range, and losses are consequently higher than would be expected in a conventional design.

The outline concept was based on various crystal ladder filters described in *RadCom* at various times; by measuring the parameters of a single crystal it was possible to choose a suitable value for the terminating impedance to provide acceptable loss and bandwidth figures. Capacitor

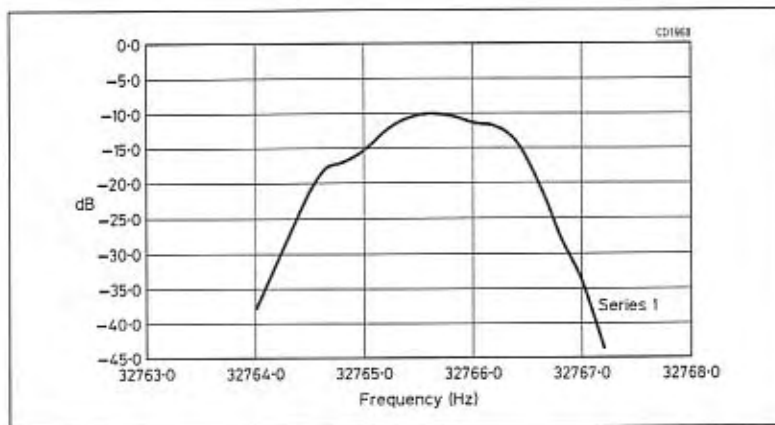


Fig 2.14: Frequency response of the 5-pole 'watch crystal' filter.

values were obtained by scaling for frequency and impedance from the published values. A value of $56\text{k}\Omega$ was chosen for R_0 and a breadboard filter constructed to evaluate the idea. The frequency response of the five-pole design is shown in Fig 2.14 from which it can be seen that it is far from optimum, and there is considerable scope for fine tuning the design to steepen skirt bandwidths perhaps, and reduce ringing. However, for very narrow bandwidth reception where the signal bandwidth is considerably narrower than 1Hz the filter response is not too important, rejection of close in interference being its main function. These crystals are very low cost and it is worth perhaps purchasing 10 or 20 at a time in order to select the best ones for an optimised filter. However, any five identical types should work adequately.

IF gain is provided by a FET buffer, followed by operational amplifiers. Manual gain control is provided simply by varying the feedback of these stages. No IF amplifier details are shown as one has not been tested, but at 32kHz there should be no problems with modern op-amps. Low noise devices such as the NE5532 should be used here.

The second mixer and BFO presented the greatest problems with regard to locking to a standard reference. The peculiar value of IF dictated by the filtering means that a BFO frequency in the region of 32.763kHz would have been needed to convert directly to a baseband in the 1-2 Hz region. A second DDS module could have been used here, but this was considered a little extravagant for generating a single tone.

Instead, the IF is mixed to an audio frequency of 765.5Hz by a 32,000Hz BFO. This is generated by direct division to 8kHz from the 5MHz reference, and multiplication. Instead of employing a separate $\times 4$ multiplier followed by a conventional mixer, a sampling mixer is used. This multiplies the incoming signal with a narrow spike at a repetition rate of 8kHz – effectively combining the multiplier and mixing functions into one. The sampling mixer is a single-ended design consisting of a transistor configured as a straightforward amplifier, switched on by short pulses at the 8kHz rate. The 765.5Hz output is also available for feeding directly to software packages such as Spectran.

We now use the fact that the Nyquist law on sampling frequencies and aliasing defines the *bandwidth* of the signal as the determining frequency, not its centre frequency. As the bandwidth of the audio tone is less than 3Hz this can be applied directly to an analogue-to-digital converter which is triggered at 12.8Hz sampling rate. This has the effect of performing a further effective downconversion by mixing with the 60^{th} harmonic of 12.8Hz (768Hz) to

give an apparent baseband centred on 2.5Hz. The 768Hz component never appears in practice, but is an artefact of undersampling a signal at frequencies well below the Nyquist rate. The only criterion is that the A/D converter must be capable of operating at the audio frequency, but at 765Hz this is hardly an issue. The 12.8Hz sampling signal is generated by dividing the 8kHz BFO signal by a further 625 times, giving a completely frequency locked system. The required resolution of the A/D converter is not as high as may be thought. For conventional wideband DSP systems, the audio feeding the soundcard (or whatever) may have a bandwidth of several 100Hz and contain adjacent transmissions and QRM. The number of bits, or its resolution, has to reflect this and 16 bit systems are the norm, giving 80-90dB dynamic range. However, here most of the signal processing has been done by analogue means and since the whole point is to look for signals buried in noise, a much lower dynamic range is acceptable.

If we assume a value for the narrowest post-processing bandwidth that is likely ever to be used, it becomes possible to estimate the dynamic range and hence number of bits needed for digitisation.

Quantisation noise (or the errors introduced by constraining the signal to fixed binary levels, coupled with the maximum input amplitude for the A/D converter) means that there is 6dB of dynamic range available for each bit of the binary word after digitisation, less a 'fiddle factor' usually taken to be around 2dB. Assuming a signal detection limit in $200\mu\text{Hz}$ (yes, micro Hz ! which means monitoring for a couple of hours at least to determine signal presence), this reduces the noise floor from the 1Hz analogue filtered value by a factor of 5000, or 37dB. Allowing for a S/N ratio of 20dB in this bandwidth – anything higher could make use of a faster processed bandwidth anyway – means that a dynamic range of $37 + 20\text{dB}$ is the minimum needed. Furthermore, a rule of thumb for digitising noisy signals says that quantisation noise should generally be 10dB less than the system noise so we are now looking at an A/D converter dynamic range requirement of 67dB. This fits in very well with commonly available 12-bit devices which can achieve close to $(12 \times 6) - 2 = 70\text{dB}$ dynamic range. For more conventional processed bandwidths of tens of MHz a 10-bit or even 8-bit A/D converter may prove adequate.

Receiver control can conveniently be handled via a single PIC controller. This can simultaneously control the DDS from the PC (or even from a rotary encoder), read a low cost serial 12-bit A/D converter, then format the data and send to the PC via the serial link. At the modest rate of 12.8 samples per second, even a 300 baud RS232 link is adequate for this, allowing for two bytes per data sample. By using the 5MHz reference to provide the PIC clock, a buffered TTL level signal is available to directly drive the 8kHz divider and the DDS chip itself. The elimination of a separate clock oscillator also reduces the possibility of spurious signals being generated. In this case, the 12.8Hz can be generated by PIC software itself rather than by employing a separate divider.

Sensitive Loop Antenna for 136kHz

By Tony Preedy C Eng, FIEE, G3LNP

The 136kHz band has proved itself to be capable, even with our low power limitation, of providing the greatest surface-wave daytime range of any amateur frequency allocation. However, the band does present a few challenges. Apart from the problems encountered in achieving 1W ERP from a typical antenna having less than 0.05% efficiency, the main challenge provided by the 136kHz band is reception in the presence of both man-made noise and powerful commercial and broadcast transmissions, some of the latter being only 15kHz from our allocation. For the majority of us, who live in rural locations with inevitable QRM from domestic mains driven appliances and lacking the space for 2km Beverage antennas, the antenna that will give us the best chance of consistently hearing DX signals in this band is likely to be a strategically sited magnetic loop.

Sources of LF QRM

The types of QRM that we are most likely to encounter

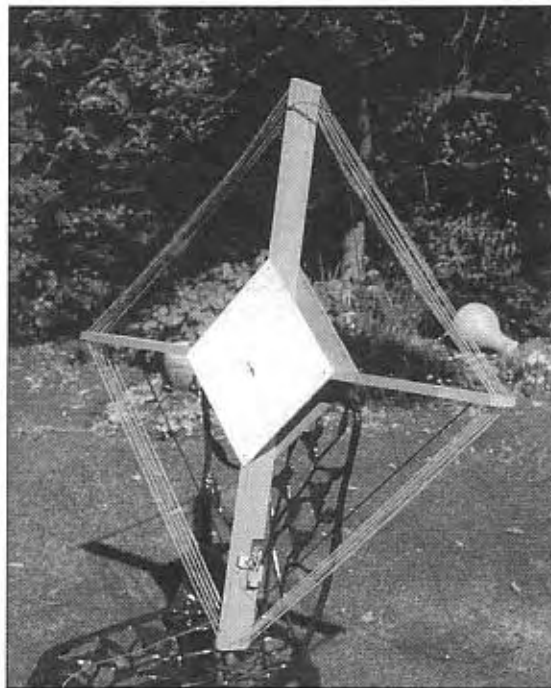
1. Continuous crackle, due to mains equipment such as fluorescent lights, motor commutators, faulty power line insulators, etc.
2. Occasional bursts of crackle, due to thermostats and solid state power switches.
3. Spurious signals due to finite dynamic range within the receiver system (broadcast IPs being the most common).
4. Static crashes due to local electrical storms (worst in the summer).
5. Unwanted sideband components from signals nominally outside our band.
6. Harmonics of TV line oscillators.

If the location is very quiet, such as I have found in the uninhabited parts of North Wales, we may be able to hear more persistent atmospheric noise, as distinct from occasional local static crashes. This is what we should be aiming for as the limiting factor to our reception capability.

Receiver Limitations

The noise blankers normally fitted to amateur receivers are not effective in reducing any of the above effects at LF, because they are wide-band devices and cannot differentiate between noise in our pass-band and the adjacent strong signals. A peak clipping device between radio and headphones or old fashioned noise limiter is effective in reducing the harmful effects to one's hearing of static crashes, although it may not improve readability.

A directional loop antenna will help to eliminate the effects of any one and (if we are lucky, more than one) of the above sources of interference. Many of the receivers, including transceivers with a general coverage capability, currently available are deliberately insensitive in the LF region, where their designers presumably expected us to listen only to strong broadcast transmissions. Unlike receivers of earlier generations, they tend to rely on simple low pass filters feeding directly to mixer stages, and consequently lack the front-end selectivity that will effectively extend the dynamic range sufficiently for our purpose. Because of this economy they inevitably have no



A simple timber frame supports the main and notching loops, plus the pre-amp.

RF stage and consequently are insensitive at 136kHz. When I first transmitted on this band I asked friends to listen for my signal. Several, even though they were no more than 200km away, could not hear me when they used a combination of 80m dipoles and the general coverage ranges of their radios. Most were hampered by lacking a narrow filter and/or having a high noise level. When I received a 579 report from a station on the north coast of Scotland, several hundred kilometres away, we realised how insensitive the common receiving set-up was at 136kHz!

A tuned loop antenna can provide the required front end selectivity necessary to reduce the effects of '3' above, whilst incorporated pre-amplification will overcome the lack of receiver sensitivity.

The loop antenna described below was required to be installed remote from mine and neighbours houses, hence away from the usual sources of QRM. In my case the antenna was placed in a timber framed greenhouse at the bottom of the garden. To minimise strong signal overload effects it was decided to have only sufficient bandwidth to receive the amateur allocation and thereby also eliminate any need for tuning adjustment after installation.

The loop was to be balanced to minimise pickup of local electric field interference. A single length of inexpensive coaxial cable was required to carry both the signal and supply voltages between receiver and antenna.

Apart from the intermittent QRM of '2' and '4' above, most receiving situations have one predominant offending signal or local noise source, the impact of which can be significantly reduced by turning the loop for minimum pick-up from this particular direction.

The ability to subsequently rotate the loop was not considered to be important, because to do so was not expected to improve the audibility of signals received from the direction of the interference source. If I have to receive

from the null direction I use the transmitting antenna and accept the extra noise that it introduces.

Loop Formula

If the required resonant open circuit loop voltage is 'E', when the loop is orientated for maximum output from a field 'e', the loop sensitivity is the ratio 'E/e' given by:

$$E/e = 0.00002 \times N \times A \times Q \times f$$

Where: N is number of turns

A is loop area in square m

Q is frequency/bandwidth

f is frequency in kHz

Antenna selectivity depends on the Q of the loop winding, therefore we need to know the Q necessary to achieve the required bandwidth.

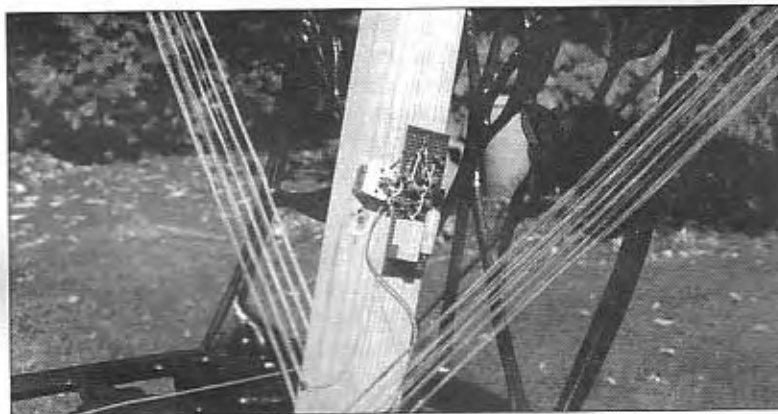
For our allocation of 2.1kHz, centred on 136.75kHz, we need a Q of $136.75 / 2.1 = 65$. The loop designs of VK5BA in *Amateur Radio* for August and September 1990 were studied. From these it was apparent that to obtain the required Q, an unscreened loop on a frame of 0.8m side would be required. The loop area 'A' was therefore fixed at 0.64m². A comparable Q can easily be achieved from a ferrite rod antenna, but in my experience amplifier noise then becomes significant at the sensitivity required for weak signal reception.

My first attempt at winding the loop used enamelled wire of 0.3 mm diameter. This gave a measured Q, with the amplifier connected, of approximately 40. A second attempt used Litz wire of the same overall diameter. This gave a Q of approximately 70, which was acceptable. By extrapolation, the optimum solid wire size, to give a Q of 65, is about 0.7 mm. Thicker wire may require the addition of a high value shunt resistor to reduce the Q.

How Many Turns?

Now we need to see how many turns the loop should have if it is to provide adequate sensitivity under practical reception conditions. Theoretically there is no limit to how much gain one can incorporate into an active antenna. One obvious possibility is to emulate a quarter wave monopole. The sensitivity E/e of a loss free monopole is such that even under the quietest conditions it would deliver over 1000µV of noise. Would you believe a report of 'S9 plus 30dB but you're only just equal to the noise'? How would your receiver cope with the accompanying 12 volts or so from LF broadcasters? In order to be able to give meaningful reports to other stations, the loop should have the same sensitivity as the associated transmitting antenna. My top-loaded 20m vertical radiator requires 18dB of pre-amplification at the quietest time in order to produce a noise reading of 5 on the S meter of my radio

Detail of the pre-amp, which is fixed to the upright supporting member of the antenna.



when using a 500Hz filter. This level of pre-amplification may appear excessive, but when a narrow IF or DSP filter is added for CW reception you will wonder where the noise went! The equivalent noise voltage was calculated to be: $E = 12.5\mu\text{V RMS}$ into the pre-amp.

Table 1 shows that at 136kHz the equivalent mean UK winter atmospheric noise field strength 'e' on a quiet day is 2µV/m in a bandwidth of 500Hz. This can sometimes be experienced in the morning at mid winter although in most situations, with the exception of local static crashes, the predominant noise is more likely to be man-made. Therefore our required loop sensitivity of E/e is $12.5 / 2 = 6.25$. This is almost 40dB less sensitive than the monopole above!

Incidentally, Table 1 illustrates both how pointless it would be to attempt to make contact over any distance on LF in the tropical zone, and why there are no LF broadcasters in that region. Rearranging the sensitivity formula we can obtain the number of turns:

$$\begin{aligned} N &= (E/e) / (0.00002 \times A \times Q \times f) \\ &= 6.25 / (0.00002 \times 64 \times 65 \times 136) \\ &= 55 \text{ turns} \end{aligned}$$

There is a practical limit, because the capacitance between turns will make the loop resonant below the required frequency if too many are used in an effort to increase sensitivity. With this number and a sectionalised winding it is still possible to resonate the loop with a small amount of external capacitance. 54 turns were used because an even number places the centre tap close to the ends of the winding.

Should the Loop be Screened?

The conventional answer is yes, because a screened loop will help to eliminate pick-up from local QRM. The argument is that local interference has predominantly an electric field, to which a screened loop is insensitive. However, the electric and magnetic fields are comparable after moving an eighth of a wavelength from the source.

The practical answer is no, because a screen increases self-capacitance, with consequent Q reduction. In fact a loop this size with 54 turns would be self-resonant way below 136kHz, because of the added self-capacitance due to a screen. Therefore, if we use a screened loop we have to both reduce the number of turns and accept a low Q, thus failing to achieve both required selectivity and sensitivity. Popular designs using multistrand screened or coaxial cables will generally have insufficient Q to adequately differentiate against LF broadcast signals when tuned to 136kHz. Although the lower sensitivity can be offset with additional amplification it must be remembered that the broadcast signals will also be amplified, possibly to the point where the radio is overdriven.

Provided we use a balanced amplifier we can tolerate some electric field pick-up, because it appears at both input terminals as a parallel mode voltage and is eliminated at the balun before driving the coaxial feeder. It must be obvious that electromagnetic interference is less of a problem for any loop antenna installed away from all sources of interference. However, VK5BR shows that if the loop can be located only 0.005 of a wavelength, or about 10m at 136kHz, from a source of QRM, the electric field is then only 2.7dB stronger than the magnetic field. The extent of protection required against in-band noise, either from a screen or balanced construction, is therefore quite moderate at reasonable separation distances. Another question is: should we make the loop balanced

by having a centre-tap at the loop or at its tuning capacitor?

If we use the capacitor to provide a centre tap, it will help to decouple those out-of-band parallel mode voltages at frequencies much higher than those to which the loop is tuned. However, we are often concerned with electric fields at the supply mains frequency and its harmonics. These can provide a parallel mode voltage, by virtue of the loop acting as an untuned capacitive probe, which will overdrive the amplifier to the point of non-linearity.

At 50Hz an eighth of a wavelength is 75cm and rather more than the length of most gardens! These low frequency signals can, however, be effectively attenuated by grounding the centre of the loop (we could ground both the loop and capacitor centres, but we risk forming a shorted turn if neither is truly balanced). The other practical advantages of grounding the loop centre are:

- that we only need an inexpensive compression trimmer for tuning, instead of a two-gang variable capacitor
- we do not require gate return resistors, which would inevitably load the loop.

Strong Unwanted Signals

I imagine that VK5BR, like our friends in America, does not have to contend with the effect of the LF broadcast transmissions that we in Europe cannot avoid. These can produce fields of $20,000\mu\text{V/m}$, even when 100km from the transmitter. If tuned to the broadcaster's frequency, the peak voltage for 100% modulation at the loop terminals would be: $6.25 \times 20,000 \times 2.828 = 350,000\mu\text{V}$. This is 89dB above the noise floor! Almost certainly any simple loop amplifier with only a single tuned circuit will not have sufficient dynamic range to prevent intermodulation or cross modulation problems when subjected to this sort of field strength.

My solution is a second loop, coupled to the main winding, but tuned to notch out the local broadcast signal. Experimentally I found that an auxiliary loop, using both half the number of turns and 65% of the area of the main loop, will provide 26dB of additional attenuation without significantly degrading either the Q or sensitivity of the main loop. My home is about 100km from the Droitwich 198kHz transmitter and I have had no IP problems since adding the second loop. Previously I did find many spurious signals in the band, owing to beats between Droitwich and other LF transmissions. The worst IP was that on 138kHz, being the difference frequency between Droitwich and the 60kHz Rugby transmitter. The resultant BBC Radio 4 modulation sidebands extended throughout our band. My only solutions, prior to adding the second loop, were to either use a valve pre-amplifier or fix the null in the direction of Droitwich. The former was not practicable with a single interfacing conductor and the latter was not acceptable because it was my misfortune to find the bearing of my local noise source to be at right angles to that of Droitwich! Where the frequency of an unwanted strong signal is below our band it will be necessary to either add extra turns or some fixed capacitance across the tuning capacitor of the auxiliary loop.

Loop Construction

Fig 2.15 shows how to build the loop using 60mm x 15mm planed timber for the frame and 3mm plywood for end checks. The slots for the main windings are formed by drilling equally spaced 5mm holes and making saw cuts. There are 10 turns in the centre and 11 turns in each

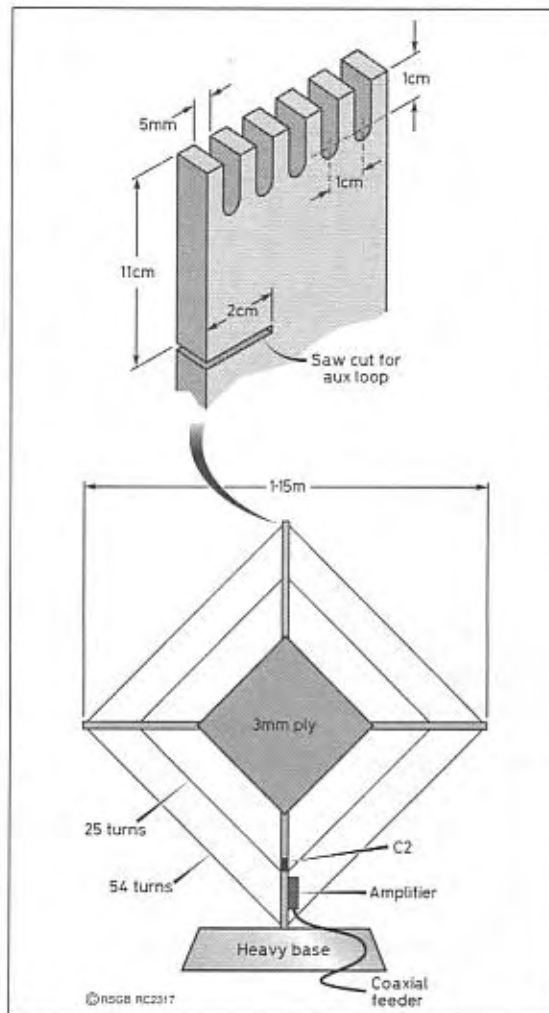


Fig 2.15: Loop construction.

other slot, with a tap at 27 turns. The auxiliary winding has 25 turns, side by side, in slots sawn into the sides of the frame. Trap tuning capacitor C2 is fixed to the frame with a small bracket of sheet aluminium. The finished loop should be varnished and mounted on a heavy timber plinth.

Pre-amplifier

In order to drive a coaxial cable with the loop output voltage, we shall need a voltage follower amplifier. The ideal input device is a JFET, because its very high input impedance will not degrade the Q of the loop. The ideal output device is an emitter follower because of its very low output impedance. As already indicated, the lack of an RF amplifier at the radio required 18dB of voltage amplification. A balanced amplifier was chosen because it offers the prospect of superior dynamic range. For the same reason, the FETs should be operated with high drain current by not biasing their gates. My amplifier does not claim to offer the best dynamic performance and I am sure this aspect could easily be improved. If a FET op-amp suitable for a single supply rail is available this could also be used. In that case the balun will not be required.

Transmitter Protection

Shunt connected, back-to-back switching diodes, plus series resistors at the gates of the input stage, allowed the loop to be placed 10m from my transmitting antenna, with its 5 amps feed current, without risk of burning out the pre-amplifier or receiver. If you do not intend to transmit on this LF band, these components can be omitted.

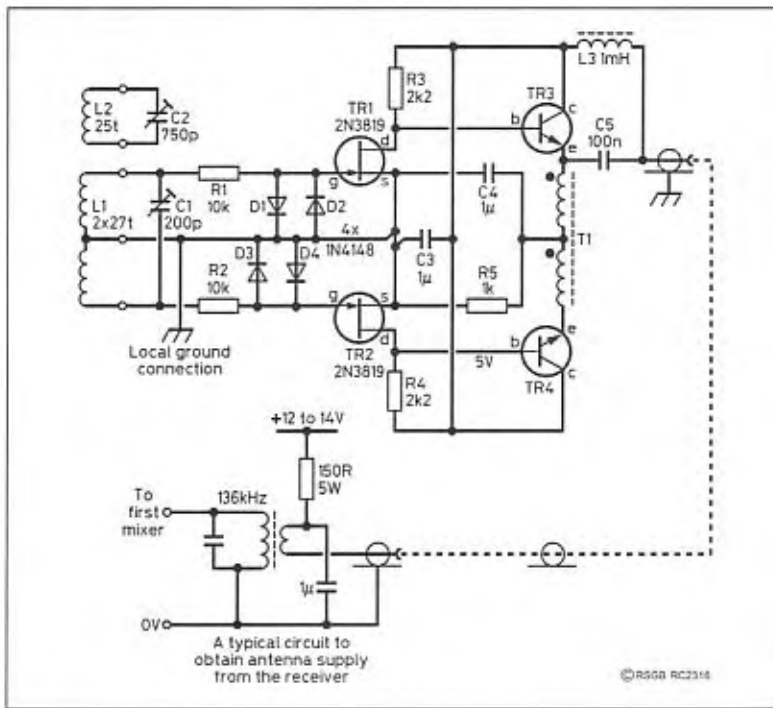


Fig 2.16: Circuit of the balanced loop and amplifier.

Amplifier Construction

I built the amplifier shown in Fig 2.16 on a piece of Vero board and fixed it to the frame of the loop, close to where the windings terminate. The design uses a minimum of components. The 2N3819s, TR1 and TR2, were matched for I_{dss} , such that the voltage at each drain is as near as possible to 5. The output transistors, TR3 and TR4, can be almost any small signal silicon PNP type. The bifilar balun T1 and choke L3 are wound on two-hole ferrite cores, 1cm square and 5mm thick, each with windings of 20 turns of fine wire. The inductance of these is not critical, because the impedance is very low at this point.

I show one method of supplying 12V from the receiver to the antenna, but other methods can be used to suit.

Adjustment

This is quite straightforward. Test the loop close to the radio first. Tune the receiver to your local LF broadcast transmission and adjust the trap capacitor C2 for minimum signal. Now tune to 136.75kHz and peak the noise by adjusting the main loop capacitor C1. Take the loop to its final position, making sure you have a good local earth connection for the screen of the feeder and the loop. Find the quietest orientation. I used headphones and a long cable from the receiver for this adjustment.

Fig 2.17: The influence of season, time, media and bandwidth on the range for manual Morse at 136kHz in the temperate zone. One watt ERP for 6dB signal+noise: noise.

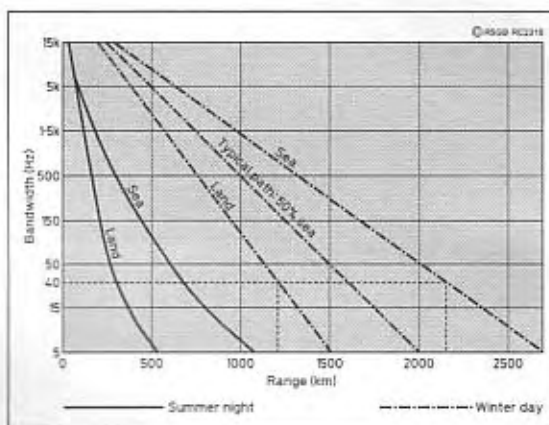


TABLE 1
Equivalent atmospheric noise field at 136kHz in a bandwidth of 500Hz.

Temperate winter day:	2µV/m
Temperate summer night:	25µV/m
Tropical summer night:	350µV/m

Results

Apart from the reduction in broadcast interference mentioned earlier, I can confirm that moving the loop 30m from the radio room and finding the best orientation reduced the noise level at the receiver by four S points, (about 24dB). Even at this distance I can detect noise from my PC monitor and lights in the house.

Fig 2.17 is based on standard surface-wave propagation curves for 136kHz, combined with the noise characteristics of Table 1. It shows how the groundwave range achievable with 1W ERP in the temperate zone is influenced by receiver bandwidth, season, time of day and path conductivity. A 6dB signal plus (atmospheric) noise-to-noise ratio is assumed. To a first approximation the range for any path can be determined by interpolating the percentage sea between the appropriate pair of land and sea curves. Ionospheric propagation should theoretically extend the range at night. However, in practice, although the commercial signals within or close to our allocation are enhanced, the disproportionately increased noise tends to mask the weaker signals of distant amateurs at night.

I gather that many users of the band also suffer other forms of reception limitation at night. In my situation it is currently the assortment of spurious modulation sidebands of the 138.82kHz signal, even though the received carrier does not significantly change in amplitude after sunset.

My IF filter bandwidth is 40Hz for CW reception, implying from Fig 2.17 a day-time range of 1200 to 2100km, depending how much of the path is over the sea. Within a radius of 2000km from my home I counted 58 country prefixes, but few as yet with an LF allocation. Some administrations have not imposed a limit on ERP and instead have stipulated maximum transmitter power. Russia for example allows 100W. Perhaps we shall hear some strong signals from there although, in the hope of being proved wrong, I fear we will be fortunate to make contact because, all else being equal, the range of two-way QSOs will inevitably be determined by the station with the lowest ERP.

COMPONENTS

Resistors (all 1/8 watt)

- R1, 2 10k
- R3, 4 2.2k
- R5 1k

Capacitors

- C1 200p compression trimmer
- C2 750p compression trimmer
- C3, 4 0.1µ mixed dielectric, 30V
- C5 0.1µ ceramic, 30V

Inductors

- L1, 2, 3 see text
- T1 see text

Semiconductors

- D1-4 1N4148
- TR1, 2 2N3819 (see text)
- TR2, 3 see text

3

Transmitters

Crystal Mixer Exciters for 137kHz

By Peter Dodd G3LDO

With all my early experimental work on LF I used a signal generator of one type or another. I did build a VFO and although it worked, the Ls and Cs were rather large and altogether not a practical proposition. Most methods used by the LF group involve synthesizers or higher frequency VFOs and digital dividers or transverters, which judging by their performance on the LF band seem to work very well. However, although I can make simple digital circuits work, all my attempts to construct any sort of VFO/divider or synthesizer have ended in failure. I am not sure of the cause of such a reliable failure rate with more complex digital circuits – I strongly suspect incompetence.

It is not the case that all my amplifiers oscillate and my oscillators don't. I seem to be able to make ugly construction RF and AF circuits that work first time so I was looking a solution that did not involve digital circuits. I remembered seeing an article in 'Technical Topics' in *Radio* on a VXO mixer. This used two crystals with one active device, which seemed rather parsimonious so I decided to make two separate VXO oscillators and put them in one box – and see what evolved.

From my crystals box I found two crystals on 10489.37 and 10355.0kHz respectively, which gives a frequency separation of 134.37kHz. I built two FET Colpitts oscillators, first with a variable capacitor across the lower frequency oscillator crystal but then found that it worked best with the variable capacitor in series with the higher frequency oscillator crystal; this gave a more constant output across the frequency range.

I did persevere with the parallel tuned oscillator. The reason was that I might need to move a crystal oscillator down in frequency. I tried using lower values of feedback capacitors so that the parallel tuning capacitor would give a wider frequency swing (the circuit stops oscillating if the capacitance across the crystal is too high).

The arrangement finally arrived at is shown in Fig 3.1 where the tuning capacitor is designated C1. The small fixed capacitor C2 is to limit the frequency range. Parallel capacitance C3 can be used to preset the lower frequency oscillator if you don't have crystals that are so conveniently spaced.

Any sort of FETs seem to work. I used some 3N152s because I had a lot of them. 2N3819s also performed fine. The output is around 300mV peak-to-peak.

By using such high frequency crystals it was easy to span the band without the need of series inductors, nor-

mally found in VXOs. My circuit is made from junk box components and the measured drift was from 137.6530kHz to 137.6524kHz after 3 hours.

Whilst trying to decide on a suitable mixer/filter I saw an e-mail from Peter DF3LP who described a very similar exciter. Peter kindly sent me a copy of his circuit (which uses 5MHz crystals), which gave me enough ideas to complete the exciter.

Apparently the first circuit of this type (that I am aware of) was built by DJ1ZB. Both these circuits are described later in this chapter.

Whether this circuit is of use to you depends on getting two crystals with the right frequency separation. This is not as difficult as it might first seem, an old VHF crystal-controlled transceiver will probably have two such crystals.

Other variations of this circuit can be found in this book. The circuit by 15TGC in Chapter 1 uses 744LS00s and 74LS02s to make a complete variable and keyed exciter. A simpler circuit, using a HEF4011B, is built into

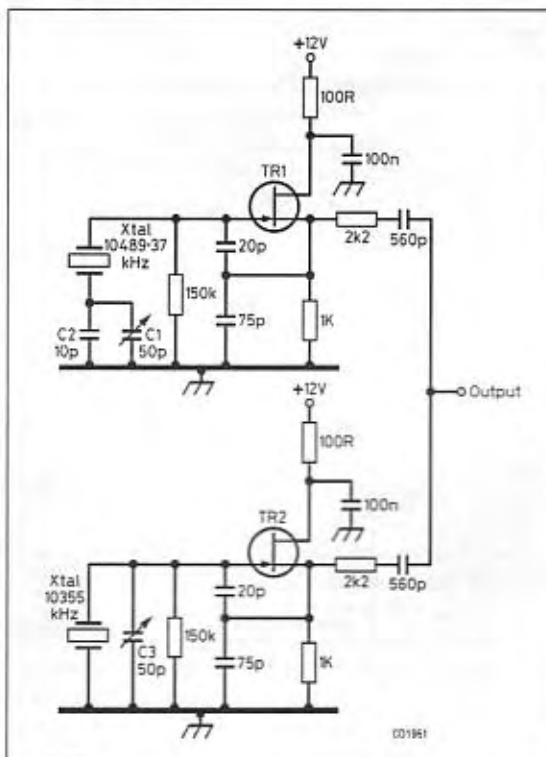
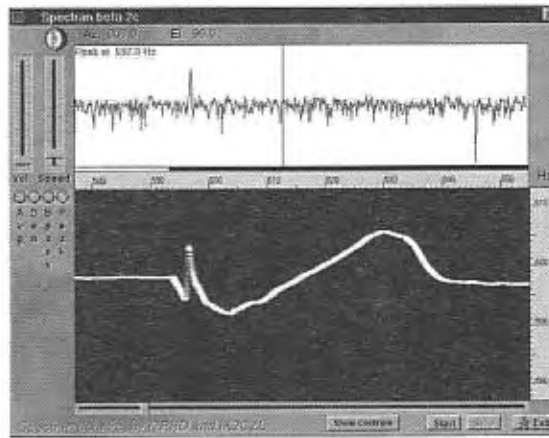


Fig 3.1: Two VXOs in one box, used in the initial testing of the crystal-mixer driver.

Fig 3.2: The G0MRF 272.6kHz mixer oscillator subjected to the hairdryer test. The frequency span of Spectran is just over 30Hz. This is a rather severe test and the compensation effect of the two oscillators can be seen.



the design of G0MRF's transmitter described later in this chapter.

One of the apparent attractions of this circuit is that if both oscillators are built in the same enclosure with identical components, the output frequency is held constant because any drift in one oscillator is compensated for by an equivalent drift in the other. G0MRF kindly lent me a

sample of his mixer/exciter to try out. What struck me about this design was its sheer simplicity. It gave a very clean sinewave out on 272.6kHz (this drives the transmitter at twice the output frequency) and was easily stable enough for QRSS operation. I gave the PCB circuit a 30 second blast with a hairdryer at one metre distance. The effect on the frequency is shown in Fig 3.2.

Alan Melia, G3NYK, sent me sets of crystals to try with the circuit shown in Fig 3.1. From these tests it is apparent the higher the frequency of the crystal the easier it is to pull in a VXO circuit (as you might expect). On the other hand higher frequency crystals are not so stable as lower ones in this circuit configuration. It is a question of compromise. The optimum frequency for crystal oscillators in this circuit for 136kHz is around 5MHz. G0MRF uses 8MHz, but his exciter is required to generate a signal at twice the transmit output frequency – and this seems optimum for his design.

The temperature variations in my garden shed shack are rather wide. I compensate for this by placing the exciter box in a large wooden instrument box. The temperature is maintained by a 15W lamp whose supply is fed via a refrigerator type thermostat.

A Transmitter for the 136kHz Band

By Hans-Joachim Brandt DJ1ZB

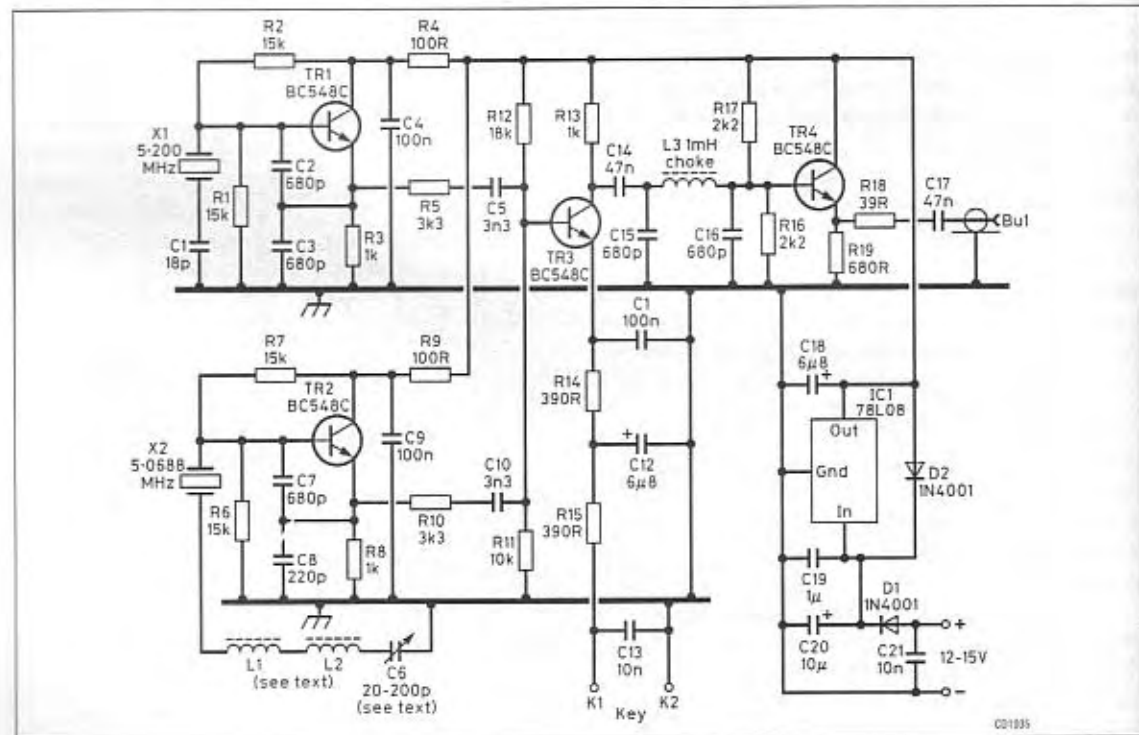
LF Crystal Mixer Exciter

This exciter produces a drive signal for the 136kHz band by mixing the outputs from two Variable Crystal Oscillators (VXOs). With the exception of the frequency pulling circuit, both crystal oscillators are identical. The circuit is shown in Fig 3.3.

The VXO is based on a design for a QRP transmitter for the HF bands. It uses available microprocessor crystals

around 5MHz. The crystals with the closest frequency difference of about 136kHz were 5.0688MHz and 5.200MHz. The 5.0688MHz crystal had often been used in QRP transmitters for the 10MHz band (on the second harmonic) and the frequency could be pulled from about 10.135MHz down to below 10.100MHz. This was achieved using a 100µH moulded choke and a 200pF capacitor in series with the crystal. It turned out to be very

Fig 3.3: Circuit diagram of the DJ1ZB LF crystal mixer exciter.



easy to pull this crystal in the range 5064.3kHz and 5062.2kHz and to mix this VXO output with 5200kHz to achieve the necessary frequency variation for the LF band.

The design of the VXO circuit depends much on the available variable capacitor, and some trials may be necessary here. I used a 20p Philips brass trimmer, which required a moulded choke of 100 μ H in series with a smaller choke, and 10p in parallel with the variable capacitor to cover the LF band.

In order to find the optimum series inductance it is a good idea to fix both moulded chokes so they lie in parallel, in close proximity on the PCB; held in place with leads long enough to allow the chokes to be moved. By varying the distance between them the coupling, and hence the total inductivity, can also be varied.

The rest of the circuit is rather simple. The output from the emitters of the two crystal oscillators is fed to a third transistor operating as a mixer. The sum of both emitter resistors has been determined by experiment for optimum mixer gain. A low-pass filter with an impedance of 1k Ω and a cut-off frequency of about 200kHz in the collector circuit removes the original frequencies, and the difference frequency is passed to an emitter follower. Because emitter followers tend to oscillate when an output cable is connected, resistor R18 (39 Ω) has been inserted to prevent this. The output voltage is about 0.8V_{eff} and sufficient to drive the power amplifier described later.

For CW the mixer emitter current is keyed. The two halves of the emitter resistor in conjunction with a 6.8 μ tantalum capacitor form an RC low-pass filter to achieve a simple method of soft keying. The value of the capacitor may be changed to alter the keying characteristics. Ensure that a capacitor is used having a relatively low leakage current, otherwise some residual output during 'key-up' (spacer) may be present.

Some remarks about the voltage stabiliser circuit are appropriate. Initially, the value of C19 was 0.33 μ , as recommended by manufacturers. The exciter worked as expected with the key down, but with the key open there was some output around 58kHz. This turned out to be LF noise generated by the IC and radiating from the DC supply cable!

The problem was fixed by replacing C19 with a 0.1 μ capacitor and 10 μ electrolytic capacitor in parallel to the input of the voltage stabiliser IC.

Power Amplifier

This power amplifier gives 100 watts out when used with a 50V supply.

In contrast to many designs employing n-channel power-FETs, this approach has been tried to avoid any problems with power transformers or chokes. The circuit of the PA is shown in Fig 3.4.

TO-3 case PA transistors BD311 (TR5 = NPN) and BD312 (TR4 = PNP) have been chosen because of their high power dissipation rating of 150 watts, equalling the total input of the amplifier. However, plastic power transistors like the BD249B and BD250B may also be tried. For optimum efficiency of the amplifier (~50%) it is important that very low RBE resistors of 3.3 Ω or even lower (2.7 Ω) are used.

Driver and pre-driver transistors are not critical but should also have a collector voltage rating equal to the supply voltage. The DC output voltage of 25 volts (one half the supply voltage) is set to the correct value by R1

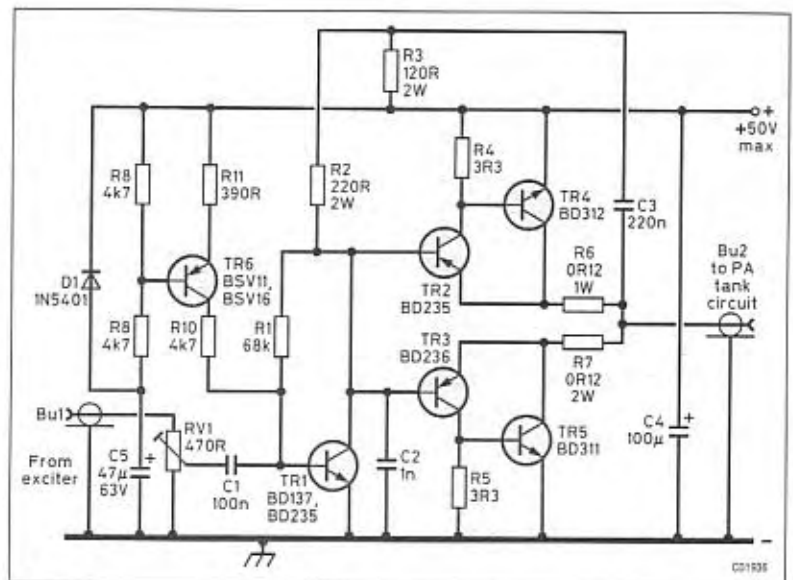


Fig 3.4: Circuit of the DJ1ZB complementary Darlington-pair power amplifier.

(dependent on the DC current gain of transistor TR1). Capacitor C1 suppresses oscillations. C3 is a bootstrap capacitor typical for this type of amplifier, providing an additional collector voltage for the pre-driver TR1 during positive PA output voltage cycles. A DC blocking capacitor at the amplifier output has not been used because one is incorporated in the tank circuit, described later.

This amplifier is driven by the crystal mixer LF exciter described above. Typical DC load current with an antenna correctly tuned and matched to 50 ohms will be about 4A at 50V supply voltage.

Owing to the conservative ratings of all components, and the two emitter resistors R6 and R7 in each of the complementary darlington group of driver and PA transistors, the amplifier will stand maltreatment, such as a detuned or even unconnected antenna.

The circuitry at the base of PNP transistor TR6 ensures a slow rise of the DC output voltage from zero to 25 volts when the supply voltage is switched on. This voltage rise charges the capacitors in the output tank circuit, and the maximum current of transistors TR2 and TR4 may be exceeded if this charging is too fast. In conventional PA amplifiers this charging of the tank circuit is done via the PA choke which is not current-sensitive.

The need for some protection circuitry became evident during the design phase after changing from a variable power supply of 6 to 30 volts (which could be adjusted smoothly) to the final fixed 50V supply. After the incorporation of this protective circuit (and the discovery of the importance of the extremely low PA transistor RBE resistors) no more transistors were lost.

The function is as follows: As soon as supply voltage is applied to the amplifier, capacitor C5 must be charged, causing a defined collector current in transistor TR6. This current is fed into the base of pre-driver transistor TR1 and saturates it, so that for a second or so its collector voltage remains very low. After C5 has charged, the collector current of TR6 will slowly decrease to zero, and the normal collector current of TR1 will flow, as set by resistor R1. The slow rise of the DC output voltage may be observed on an oscilloscope or even with a simple DC meter. When the supply voltage is switched off, capacitor C5 is discharged by diode D1, ready for the next smooth rise of the DC output voltage.

The whole PA circuitry is built on a large heat sink, using mica (or the modern plastic) washers for the isola-

tion of the transistors (except TR6). The heat sink should be grounded and connected to the negative supply voltage. Other components can be placed on mounting strips with soldering lugs and wired the old fashioned way.

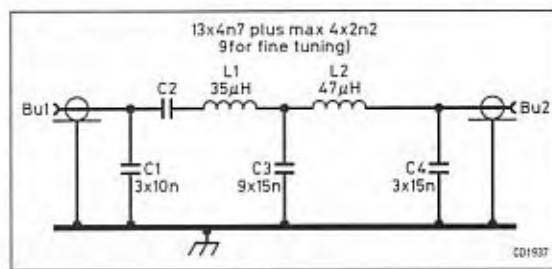
Amplifier power may be raised by employing transistors with higher voltage ratings (BD313/314 = 80V VCEo). Ensure that a higher VCEo is combined with a lower fT; >1MHz for the 80V and 100V types BD315-318 compared with >4MHz for BD311-314 types. With VC being the total supply voltage and VCrest the transistor saturation voltage (which may be neglected at this high supply voltage), the optimum load resistor can be determined by the formula:

$$R_o = \frac{(VC - VC_{rest})^2}{8 \times P_{out}}$$

Tank Circuit

The configuration of this 'peaked low-pass filter', see Fig 3.5, has evolved through many years in the design of HF

Fig 3.5: Tank and filter circuit for the DJ1ZB transmitter.



QRP transmitters. This design appears to eliminate the tendency of the solid state PA to develop parasitic oscillations when the antenna is tuned. With this output filter the input capacitor C1 is not critical with a push-pull PA and it takes care of the higher order harmonics only.

The output impedance of the PA, described earlier, is 3 ohms (at Bu1). This is transformed to 50 ohms at C3 by C2 and L1. C2 also acts as a DC block for the PA supply voltage.

C3, coil L2, and the output capacitor C4 operate as an additional low pass filter for Z = 50 ohms. Good harmonic filtering is necessary as the third harmonic of the amateur LF band will fall into the aeronautical beacon band. The tank circuit has been calculated first and then optimised using the ARRL Radio Designer Program.

In general, the permissible operating voltage of all fixed capacitors necessary for LF tank and harmonic filter circuits is frequency dependant; that means, the voltage ratings printed on the capacitors for either DC or 50Hz AC are not valid for higher frequencies. Therefore, when designing such a tank circuit the best way is to consult the manufacturer's data book for these capacitors. The reactive power is highest in the circuit C2-L1, which operates at a Q of about 6. Because of this selectivity, C2 (with a total of about 65n) is made up of 13 x 4.7n to withstand the reactive power in this circuit, and up to four smaller capacitors of 2.2n are added for optimum fine tuning to 136kHz. These small capacitors also have to tune out the tolerances of all capacitors (typically (10%)).

A Crystal/Mixer Controlled 136kHz Transmitter

by Peter Schnoor DF3LP

This small low power transmitter uses a simple VXO/mixer to produce a variable, but stable, CW signal on 136kHz. Depending of the pi-filter components used and voltage for the final stage, this transmitter should be capable of 100 watts output.

Exciter

The exciter comprises two crystal oscillators, which produce a 136kHz signal by mixing the outputs of the two oscillators. The circuit, shown in Fig 3.6, is a slightly modified version of the DJ1ZB-Exciter described earlier in this chapter. One of the oscillators is a VXO, which gives a frequency coverage of about 2.5kHz on the 136kHz band.

The exciter is inherently frequency stable because the two crystal oscillators are thermal coupled. Any frequency drift in one oscillator due to temperature, is compen-

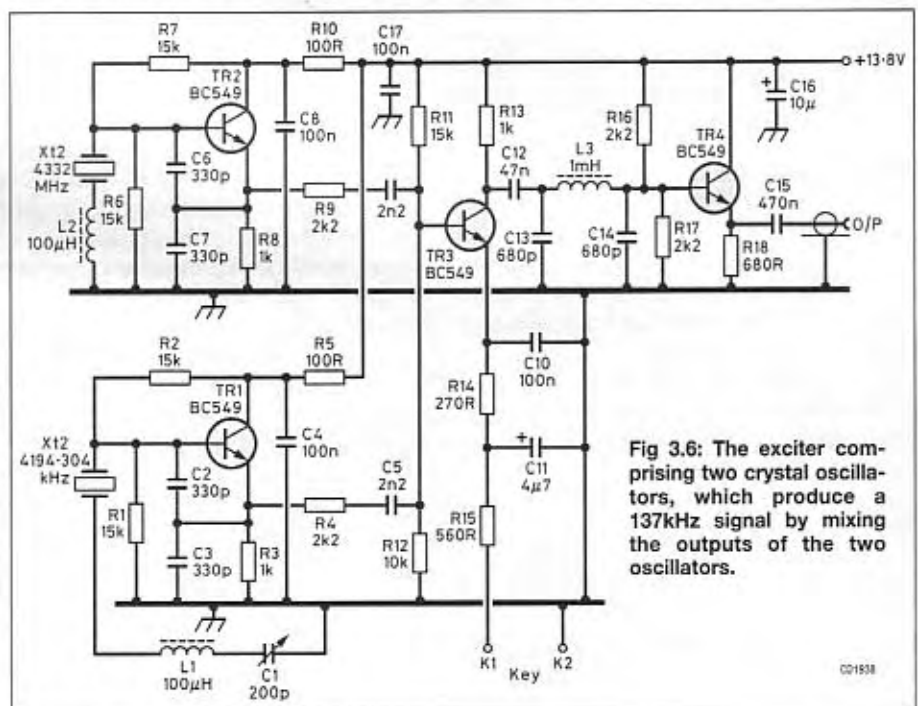


Fig 3.6: The exciter comprising two crystal oscillators, which produce a 137kHz signal by mixing the outputs of the two oscillators.

sated by the same frequency drift in the other; this is provided the construction of the oscillators is identical.

The conversion product is filtered and amplified by an emitter follower, a square wave pre-driver and the switching stage to drive the power amplifier, see Fig 3.7. This power amplifier does not make use of any output transformer since ferrite toroids for 100 - 200kHz are not always available. The prototype was built from junk box components in 2 - 3 hours.

The components of the exciter and the two driver stages are soldered directly to the copper side of a simple PC board shown in Fig 3.8. The prototype used surface mounted components; however the PC layout is designed to accommodate various sizes of components. This layout also allows modifications of the circuit if you wished to experiment with the original design.

Power Amplifier

The final stage was built separately from exciter and drivers using 'ugly construction'. The 3mH choke is wound on a ferrite toroid of unknown material, OD = 30 mm, (1.18 inches), using 1mm diameter enamelled wire, 30 turns. Running this toroid as an output transformer was not successful.

The pi-filter coil is air wound using 2.5mm enamelled wire 23.5 turns, d = 45mm (1.77in). Use only high grade polypropylene capacitors for all output filtering (WIMA, FKP-1, 1000/400 volts).

The output from the pi-filter is not a clean sinewave and an additional two stage pi-filter, 50 ohms input/output, is recommended. (Cin = 15n, L1 = 53µH, Cmid = 33n, L2 = 53µH, Cout = 15n). Amidon T200-2 cores are useful up to 500 watts without cooling but air-wound inductors (37 turns, 1mm enam. close wound over Polyacryl core, D = 50mm) are preferred. The antenna will do the rest.

An IRF630 can be used for the final stage but the IRF640 is much better since its on-resistance is lower. This small amplifier will give more power by using higher voltage and an IRF840 would be recommended.

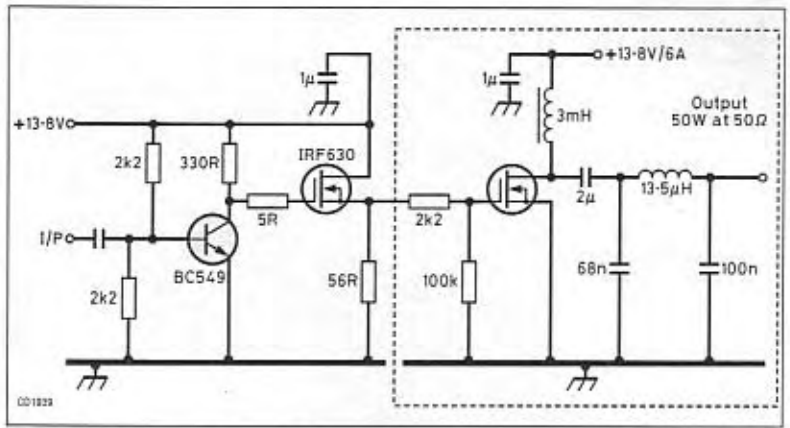


Fig 3.7: The Driver and Power Amplifier.

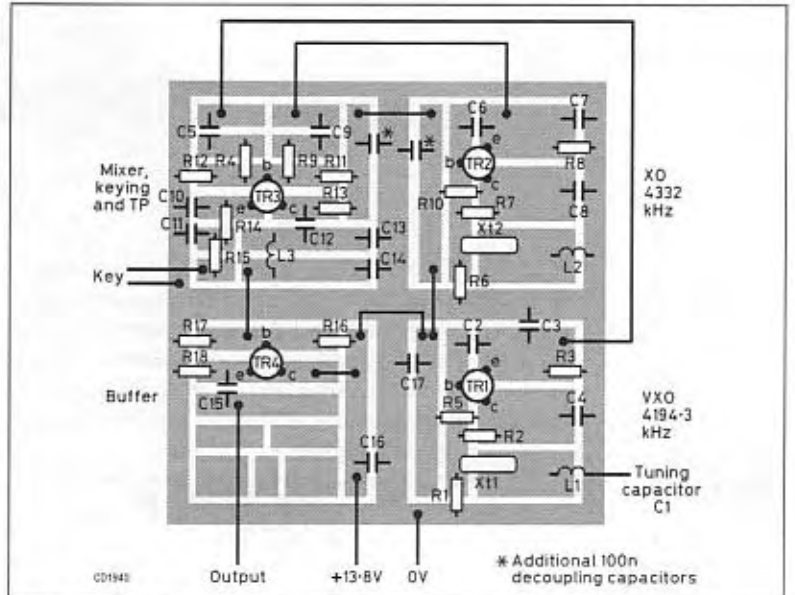


Fig 3.8: PC-Board for the exciter and driver stages.

A Transmitter for 136kHz

By Dick Rollema PA0SE

This transmitter (Fig 3.9) – uses four HEXFETs, IRF520s in push-pull parallel. The exciter comprises a HCF4060 with an 8738.89kHz crystal oscillator, set to divide by 64 to produce a drive signal at 136.545kHz. Provision has been made for a VFO to be switched in to replace the crystal oscillator. This is described later.

The output is about 120 watts with a supply voltage of 30 volts. Increasing the supply to 40 volts raises the output to about 180 watts.

The drive signal from the HCF4060 IC is a square wave, filtered by a Chebychev third order lowpass filter, with 3dB ripple, and the output increased by a class A amplifier. T1 is wound on a 23mm diameter Philips 3E2 toroid.

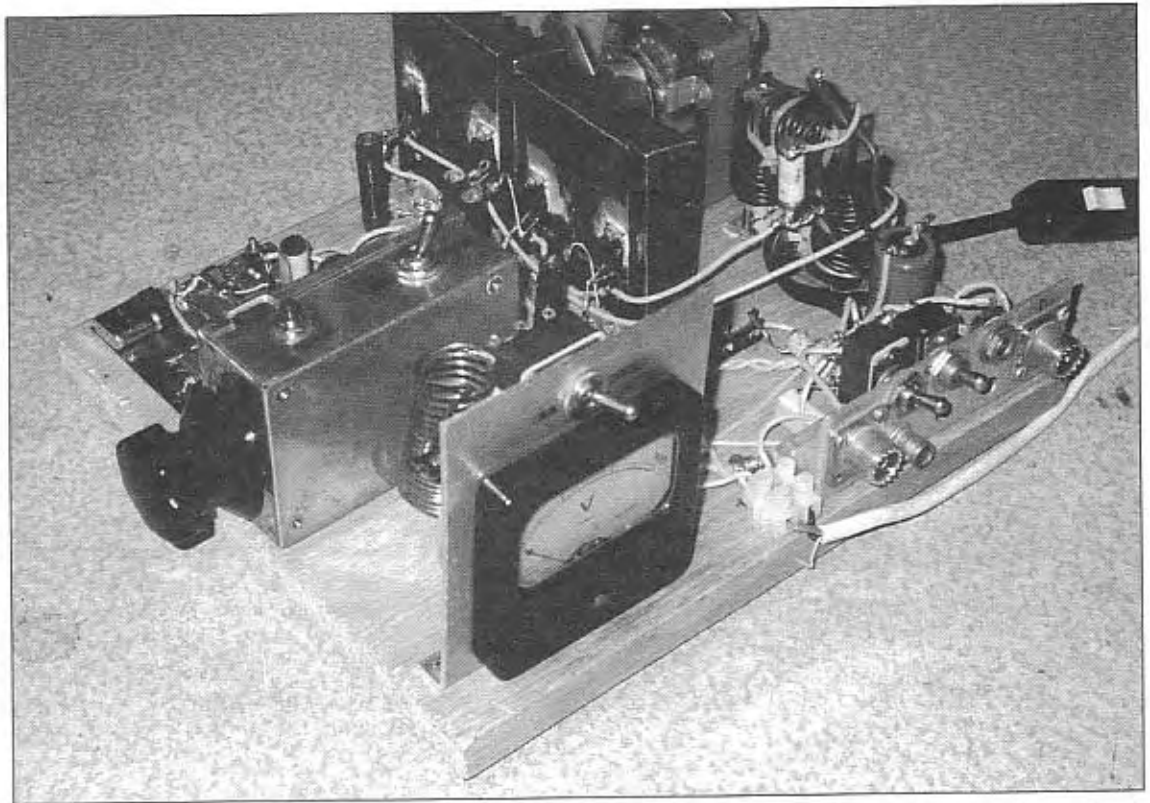
T2 is constructed – see Fig 3.10 – using the core of a TV line output transformer. The transformer windings are wound on three of the four sides of the rectangular core

using 1.5mm² copper wire with PVC isolation. (The fourth side is used for mounting.)

The secondary is first wound on the core. It comprises 15 turns of two wires in parallel in order to completely cover the three sides. The primary winding is wound over the secondary winding, one on each of the three sides of the transformer core, wound with three turns of two wires laid in parallel. The 2 x 3 turns are connected in series to 6 turns with mid tap. Finally the three primary windings are connected in parallel. This method of winding is used to make the coupling between primary and secondary as tight as possible and transforms the 8Ω output impedance of the IRF520s to the 50Ω load.

The bias to the 1RF520s is set with R7 so that they are just conducting under no-signal conditions (class B). When the amplifier output was checked with a dummy load the output was a pure sinewave when viewed on the oscilloscope. However, when the output was connected to

Fig 3.9: PA0SE's transmitter for the 136kHz band, mounted on a wooden board. On the small front panel to the right are connectors for the antenna and the receiver, the send/receive switch and the keying jack. A switch is also provided for selection of transmit antenna or active ferrite antenna for the receiver. The exciter is shown in the top left hand side and the metal box on the left contains the VFO with a push switch for netting.



a tap on the antenna loading coil high frequency oscillations were seen superimposed on the output signal, which were assumed to be weak harmonics. The HEXFETs are meant for switching and perhaps not too linear. The tap on the loading coil presents 50 ohms on 136.545kHz and a high reactance to harmonics.

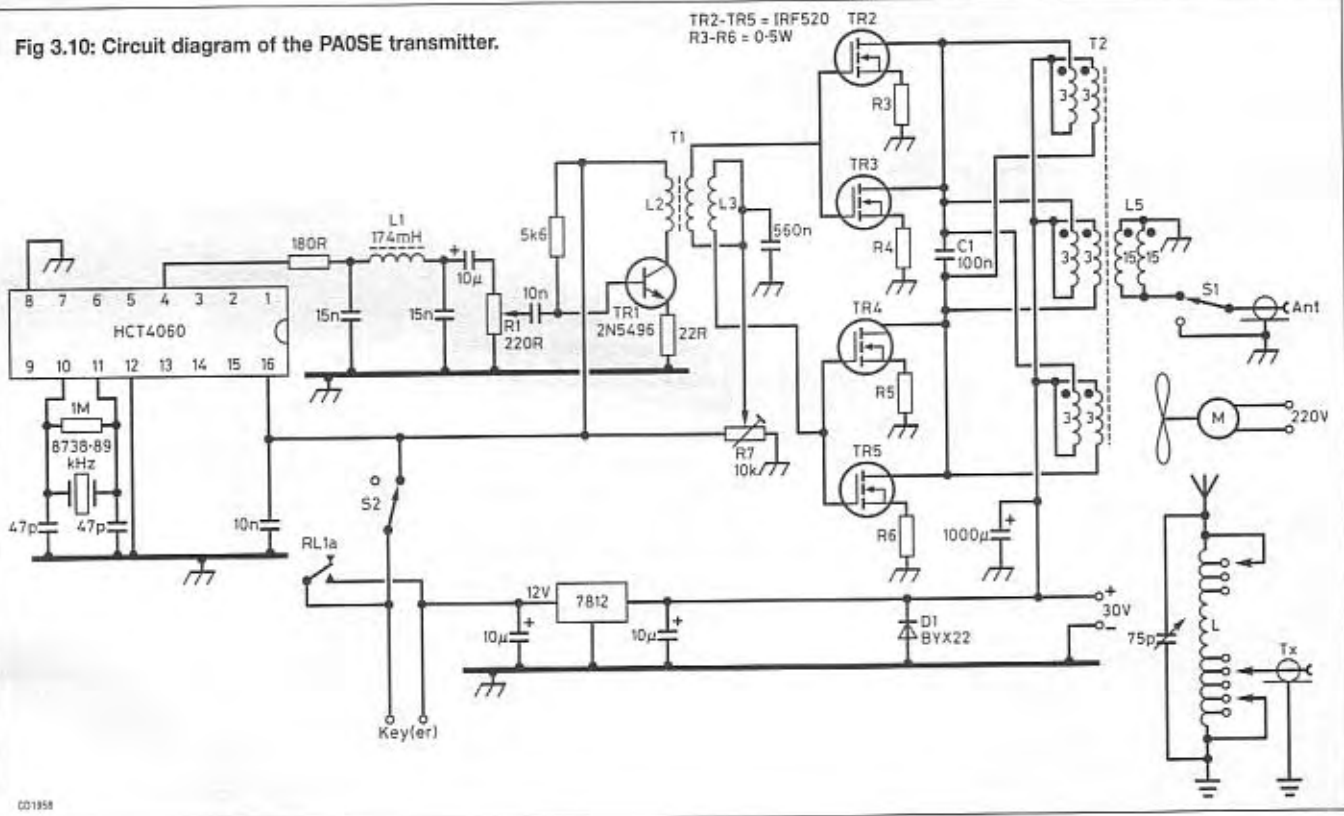
A 100n capacitor connected across the primary of TR2 (reactance $-j23.4$ ohms) eliminated these harmonics. The current through this capacitor is around 1.5A and most capacitors that I tried heated up. The exception was a

white German WW II oil-filled 50nF capacitor, and two of these were connected in parallel. See the Appendix for information on components for RF.

The supply voltage and current are monitored using a switched meter. The meter shunt is the coil at the rear of the meter shown in Fig 3.9.

The 120Ω resistors in series with the gates of the HEXFETs must be connected directly to the gates. They suppress the VHF parasitic oscillations that killed quite a few HEXFETs before the cause of their demise became

Fig 3.10: Circuit diagram of the PA0SE transmitter.



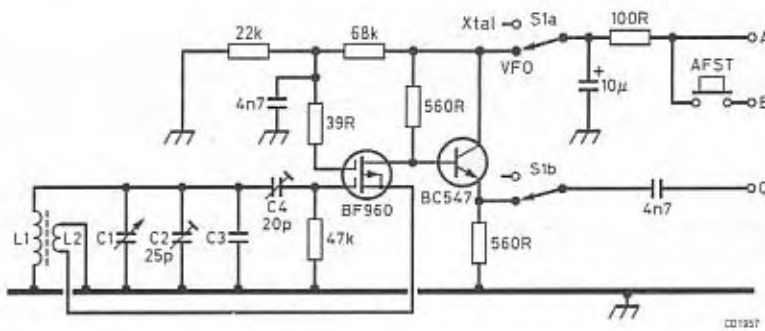


Fig 3.11: The VFO, circuit diagram. Connection points A B and C relate to the same letters in Fig 3.10.

clear. The two pairs of HEXFETs are mounted on separate heatsinks and a small fan is used to keep them cool.

The VFO

A VFO can be used instead of the 8738.89kHz crystal. The output of the VFO is simply switched in parallel with the crystal and because the output impedance is low, the crystal ceases to function and the VFO has control of the frequency. The VFO is housed in a small aluminium box as shown in Fig 3.11.

The VFO has a frequency range of 8684.8 to 8819.2kHz. This is divided by 64 in the HC 4060 to give a frequency range to the input to the PA of 135.7 to 137.8kHz.

The consequence of this approach is that the frequency stability is excellent because the small amount of VFO drift is divided by 64 on 136kHz. Coil L1 is wound on a yellow Amidon powdered iron toroid type T68-6, which

has a low temperature coefficient. The frequency range is determined by the capacitance variation of C1, which is in parallel with L1 and the capacitors C2 and C3.

$$(C2 + C3) = 32.06 \times C1 \text{ pF. } L1 = 322 \mu\text{H} / (C2 + C3).$$

In this case C1 has a range of 16.5pF. From this follows $C2 + C3 = 529 \text{ pF}$ and $L1 = 0.609 \text{ microhenries}$ (11 turns on the T68-6 core). C3 = made up of capacitors totalling about 528pF (silver mica or polystyrene type). C2 is an air trimmer of 25 pF maximum.

L2 is 1 turn wound in the same direction as L1. Starting from the earthed end, both turns go in the same direction through the centre of the toroid.

The signal level from the VFO is set by C4.

Switch S1 selects VFO or crystal control.

The TUNE push button connects the supply voltage to the HC4060 and VFO only. This results in a weak signal on the transmit frequency for zero beating on to a received signal.

A 400W Transmitter for 136kHz

By David Bowman GOMRF

This design is a complete transmitter for 137kHz. The whole circuit is contained on a single printed circuit board measuring 178 x 128mm. It can be assembled for either 300 or 400 watts output. The transmitter includes transmit/receive switching, forward and reverse power metering and in-built protection from over-current conditions and high levels of reflected power.

The circuit diagram is shown in Fig 3.13, while the general construction is shown in Figs 3.12 and 3.14

The Transmit drive is generated by a variable crystal oscillator constructed around IC1a and IC1b. The crystals are tuned differentially by varicap diodes fed by a ganged potentiometer VR1/2. The third gate, IC1c, functions as a mixer, which is followed by a simple low pass filter and a buffer amplifier. The output of IC2 is a sine wave at 274kHz.

At 137kHz, the VXO has a range of about 1.5kHz which represents over 70% of the band. Adjusting the values of C2 and C4 sets the frequencies covered.

An alternative to the VXO is to inject an external oscillator into IC3 at twice the required frequency.

IC3 divides the 274kHz from the VXO to 137kHz and provides two antiphase squarewave outputs. These drive a pair of power FETs in class D. CW keying is introduced in the supply line of FET driver IC4. Time constants around TR1 ensure controlled rise and fall times to minimise key clicks.



Fig 3.12: Transmitter viewed from the front.

The drain-to-drain impedance of the FETs is matched to 50 ohms by T2. A series of taps on the secondary of the matching transformer allow the power delivered to the antenna to be adjusted in a series of steps. After passing through the transmit/receive relay contacts the signal is filtered by a five-element low pass filter to remove the high level of odd harmonics

Protection circuits

Overcurrent protection is achieved by using a current sense resistor in the DC supply line. When 10 amps flow

00194T

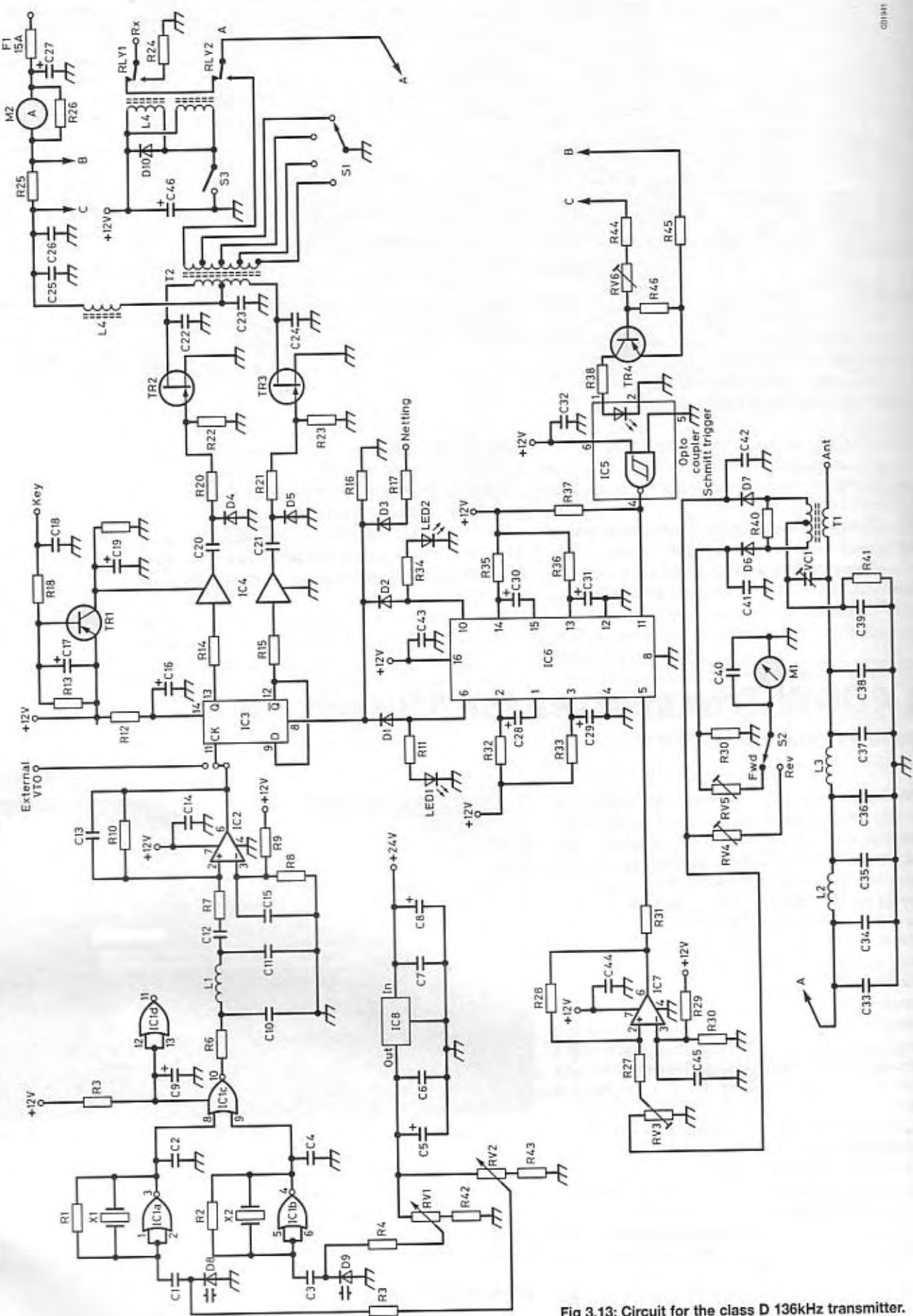


Fig 3.13: Circuit for the class D 136kHz transmitter.

through R25, a voltage of 0.7V is developed across it causing TR4 to switch on. Current then flows through R38 and the LED contained within optocoupler IC5. The output of IC5 goes low, triggering monostable IC6. The output of IC6 is connected to the Set Direct input of the 4013 by diode D2. When the SD input goes to 12V, the 137kHz drive to the FETs is inhibited. The monostable resets after a time period determined by R35 and C30.

Reflected power protection is implemented in a similar manner. Directional coupler T1 senses forward and reverse power. The voltage at the wiper of VR3 is proportional to the reflected power. When this voltage exceeds

that at the non-inverting input of IC7, the output at pin 6 goes from 12V to zero. The output from IC7 is used to trigger the other half of monostable IC6. Its output also controls the Set Direct input of IC3, diodes D1/D2/D3 acting as an OR gate.

Overall, the design is very robust, the two protection circuits responding to fault conditions in a few tens of microseconds. This ensures that the transmitter is able to withstand the frequent short circuits and high voltage arcing that can be part of everyday operating on LF.

A PCB is available from the author. Further details are available at my website – see Appendix.

COMPONENTS

Resistors

R1, R2	10M
R3, R4	22k
R5, R12, R20, R21, R45	10R
R6, R7, R31, R42, R43	1k
R8, R9, R29, R39	10k
R10, R41, R44	4k7
R11, R33, R34, R36	12k
R13	2k2
R14, R15	680R
R16	33k
R17	1k5
R18	3k9
R19	150R
R22	4k7 mounted off board
R23	4k7 mounted off board
R24	47R
R25	R07 6watt (RS)
R26	Shunt for 15A FSD
R27	47k
R28	220k
R30, R37	2k7
R32, R35, R46	100k
R38	5k6
R40	100R
VR1/2	10k dual gang pot
VR3	10k preset
VR4, VR5, VR6	22k preset

Capacitors

C1, C3	1n
C2, C4	33p
C5, C8, C14, C15	10 μ
C17, C45, C46	10 μ
C6, C7	100n
C9, C19	47 μ
C10, C11	820p
C12	470n
C13	15p
C16	4 μ tantalum
C18, C32, C41, C42	100n ceramic 50V
C20, C21	470n polyester 63V
C22, C24	10n 250V pulse capacitors 5%
C23	470n 250V polyester
C25, C26	2 μ 2 100V Polyester
C27	1000 μ 63V – not on PCB
C28, C30	22 μ tantalum 16V
C29, C31	470n electrolytic
C33, C36	2n2/1kV polypropylene
C34, C37	10n/1kV polypropylene
C35	4n7/1kV polypropylene
C36	22n/1kV polypropylene
C39	2n2 polystyrene 160V
C40, C43 C44	100n ceramic 50V

Semiconductors

IC1	HEF4011B
IC2	TL071CN
IC3	HEF4013B
IC4	TC4426CPA
IC5	H11L1 Opto isolator
IC6	HEF4538B

IC7	TL071CN
IC8	78L18AZ
D1, 2, 3	1N4148
D4, 5	MBR150
D6, 7	1N4944
D8, D9	BB405 varicaps
D10	1N4002
Tr1	BD136
Tr2, 3	STW34NB20 MOSFET
Tr4	2N5401
LED1, 2, 3, 4	LEDs Two Green for power: 12V and 40V Two Red Ultrabright. – VSWR and Current trip

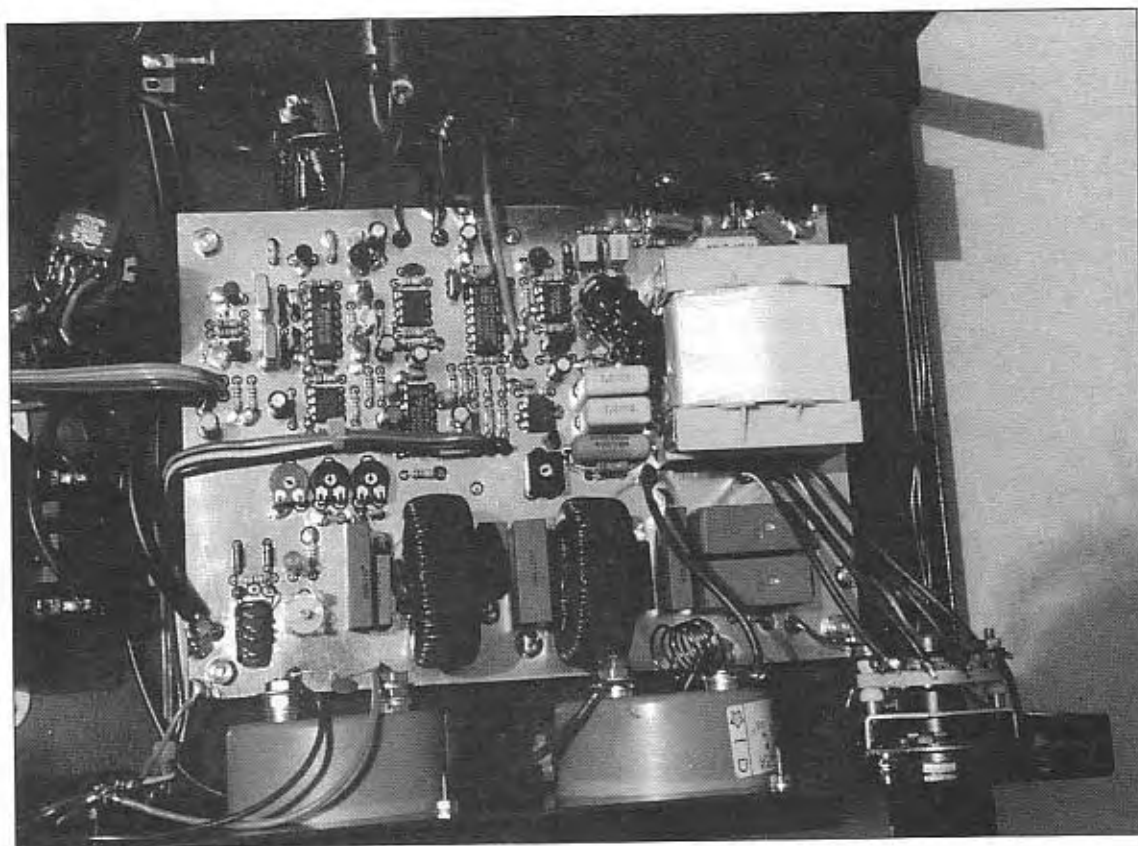
Inductors

L1	470 μ H 7BS. BEC
L2	54 μ H T157-2 powdered iron toroid. 59 turns of 0.8mm wire
L3	54 μ H T157-2 powdered iron toroid. 59 turns of 0.8mm wire
L4	11 turns 1.5mm on powdered iron toroid T92-2
T1	primary 13 turns bifilar 0.4mm. secondary 1 turn (RG58 inner) on 15mm 3C85 ferrite 42mm 3C85 ferrite toroids. 1 core for 300W. 2 stacked for 400W
T2	Primary 10 turns 1.5mm CT. Secondary 21 turns 1mm with taps at 6, 10, 15, 18 turns.

Miscellaneous

X1	8.000MHz crystal
X2	8.275MHz crystal. Quartslab. Fundamental mode, 20pF parallel load
RL1 RL2	12A relay. One pole change-over
VC1	5-55p PTFE 300V trimmer – Farnell
M1 M2	1mA FSD Meter
S1	Rotary ceramic switch 1 pole 5 way break before make
S2	Fwd/Rev switch 1 pole 2-way switch
S3	Two pole centre off toggle switch
Heat sink.	Single sided 1.2 degrees per watt
Isolating washers	TO247
Power Supply	between 30V (250W) and 45V (400W) at about 10A

Fig 3.14: Close-up of the transmitter board PCB.



136kHz 1kW Transmitter

By Dave Pick G3YXM

The transmitter described here has been developed over the last few months of operating on 136kHz and incorporates a number of refinements to the original, very simple design. Although it is a simple design it does require a reasonable level of home-brew experience to get it going without problems.

In order to radiate a useful signal on 136kHz one either needs a superb aerial and a few watts or a normal sized aerial and a few hundred watts. An 80m dipole at a sensible height, with the feeders strapped, will have a radiation efficiency of a few points of a percent, requiring 1kW or so to give the allowed 1W ERP. I therefore set out to design a Tx that is reasonably small, produces as near to 1kW as possible and will withstand a few flashovers *etc.*

I chose to use power MOSFETs in a switching design, which gives very good efficiency so easing size and weight problems.

Circuit Description

The rig is crystal or VFO controlled and uses Philips Locmos 4000 series chips to generate push-pull square-waves on 136kHz from the high frequency source. The circuit is shown in Fig 3.15.

The basic divider chain requires an input at 1.36MHz, which comes either directly from the VFO or from the 13.6MHz crystal via a further divider. The 4013 D-type, IC4, must divide by two in order to produce its push-pull output so the 4017 decade counter, IC2, is strapped to divide by five by connecting TR5 back to the reset pin.

The 136kHz square-waves are fed to the TC4426 driver chip, IC5, via 1k resistors so that undue loading of IC4 outputs does not take place when the driver chip has no supply (key up). An interlock (to prevent damage to the output devices, should the drive fail or the SWR be poor) is provided by IC3, a 4538 dual re-triggerable monostable and the gates IC1 and IC6.

IC3a is clocked by the 272kHz signal at the output of the IC2 and, if the signal stops, the output on pin 6 goes low and (via the gates) stops IC4 producing output. Should the reflected power or the PA current rise to unacceptable levels the drive is also shut off. When the error signal stops, IC3b is triggered and, by the same mechanism, prevents keying for approximately one second. This prevents the Tx chattering on and off in the presence of a fault and gives the operator enough time to let go of the key! IC3b is prevented from triggering on power-up by the R and C on pin 13 (clear).

For stability, the VFO or crystal oscillator is left running all the time that the rig is on. It causes no QRM because no 136kHz signal is generated until the divider chain is energised when switching to Tx mode. A suitable VFO for the transmitter is shown in Fig 3.16.

When the key is pressed and drive is present, TR5 is turned on, allowing 12V to be applied to the driver chip IC5, which drives the output FETs into conduction. Some may wonder why I use a complex keying system rather than just passing the drive through a gate. The answer is key-clicks. If the Tx were to be turned on and off at the

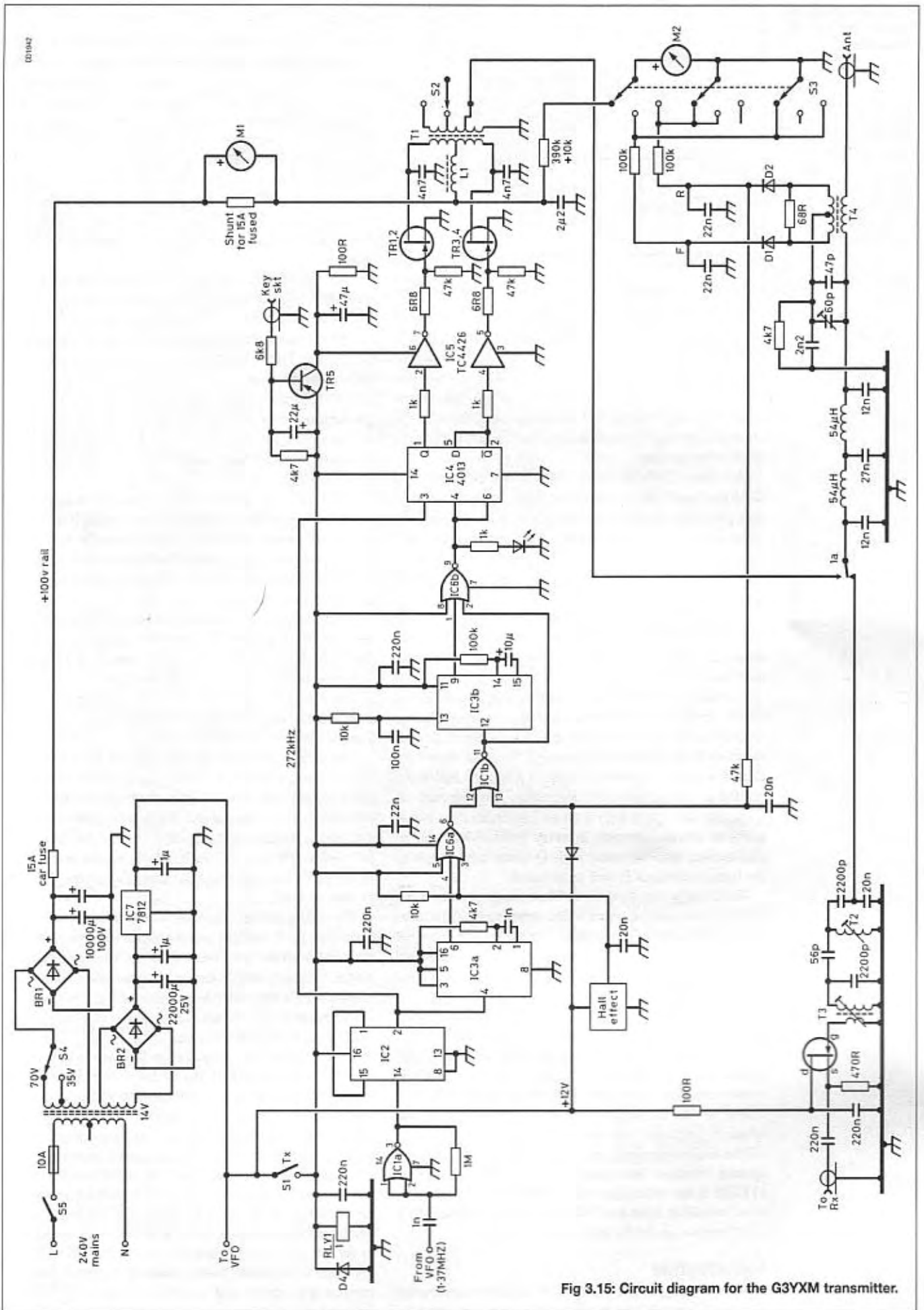
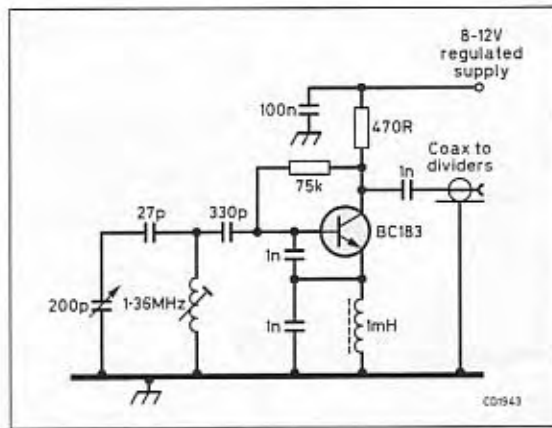


Fig 3.15: Circuit diagram for the G3YXM transmitter.

Fig 3.16: A suitable VFO for the G3YXM transmitter.



rise-time of the drive waveform the sidebands, even after the filtering provided by the high Q aerial, would extend well outside the band. TR5 turns on and off fairly slowly and the gentle rise and fall of RF is aided by the fact that the output mark space ratio of the 4426 gets less as the volts drop. Don't use its non-inverting sister, the 4427, as the mark space ratio increases on that one and the drive waveforms overlap.

The four IRFP450 power MOSFETs are operating close to class E, either hard on or turned off and so dissipate very little power and only require a small heatsink. They must be kept reasonably cool, however, as their internal resistance rises with temperature. I only show a pair of MOSFETs on the diagram but in fact each side of the push-pull pair has two MOSFETs in parallel.

The output transformer steps the voltage up by the amount chosen by the output selector switch. The more turns on the secondary, the more voltage and hence the more power into 50 ohms.

Up until now we have been dealing with square-waves which are rich in odd harmonics so an output filter is required before we dare apply this to an antenna. This is the part of the design that has given the most problems, there's a lot of harmonic energy to dissipate and some heating of components will take place. The filter roll-off frequency is set quite high at about 220kHz as there is virtually no second harmonic to worry about. When used in conjunction with the usual high Q aerial tuning system the harmonic output is well suppressed.

The SWR bridge consists of T4 and its associated components. It is wound with a bifilar winding of 2x18 turns which forms the centre-tapped secondary and the coax inner passing through the centre forms the single-turn primary. The diodes D1 and D2 must be high-speed devices capable of standing a few hundred volts. The complicated switching arrangement is to keep the loading on the bridge constant whatever the switch position.

The protection circuit cuts the drive for about a second by inhibiting keying and is triggered by poor SWR or overcurrent. The current trip uses the magnetic field generated by the choke CH1 to fire a Hall-effect device, which is positioned near one end of the ferrite rod.

The receive pre-amp uses a pair of top coupled tuned circuits giving a band-pass response over the 135 to 138kHz range with very good attenuation of the long wave broadcast band and 100kHz Loran. A single J-FET (TR8) makes up for the filter loss.

Construction

If you are going to use the VFO, it should be constructed in a small box made from pieces of double-sided copper-

clad board soldered together. Leads of components should be kept short and a bit of silicon rubber or candle wax could be applied to hold parts firmly to prevent frequency wobble. The VFO frequency will be divided by ten, so 100Hz drift at VFO frequency will only be 10Hz at output frequency. Even so, it is worth using good quality polystyrene capacitors in the VFO to counter the positive temperature coefficient of the other components. The VFO should be routed to the 4001 squarer via a short piece of coax. This gate, strapped as an amplifier, has a very high gain, so leads must be kept short and decoupling Cs applied across each chip to guard against instability.

I built the CMOS circuitry on strip-board taking care to keep the tracks short and earth all unused inputs.

The TC4426 chip IC5 is capable of driving 1.5A into the gate capacitances of the big FETs and the decoupling Cs must be fitted close to the chip with short leads. The 6R8 series resistors are actually mounted on the gate pins of the FETs, the resistor leads forming the connections to the strip board. It is probably best to use four resistors, one for each gate. The strip board should be grounded to the earth plane as near to the FETs as convenient via a short lead.

The FETs are static sensitive devices and can be destroyed by careless handling. They should be mounted so that the source leads are soldered straight down onto the ground plane. It is good idea to solder the 47k resistors from each gate to ground first, to prevent static build-up during construction.

Having applied a very small amount of heat-sink compound to both sides of the transistor insulating washers, the FETs should be clamped firmly down to the heat sink and a test made to check that the drains are not shorted to it. The two 4n7 capacitors should be connected straight from drain to source of the pairs of FETs with short wires. They must be at least 1kV pulse rated.

The output transformer (T1) should be constructed from a few inches of two-core (figure-of-eight) speaker cable wound eight times through the ferrite toroid. This is connected as a centre-tapped primary by connecting one end of one winding to the opposite end of the other. The join is the centre tap. The secondary is wound over the top of it with 20 turns of thin wire tapped at 12 and 16 turns to connect to SW2.

The wiring between the transformer and the switch will carry high peak voltages and should be well insulated and screened from the low-power circuitry. This applies to the output filter too, which should be constructed on a separate piece of copper-clad board.

The most critical components are the dust-iron cores; T200-2s may be expensive and require a lot of turns but they work better than most. To get the correct inductance you will need about 65 turns of 1mm wire. Use a small computer fan running on the transmitter +12V to keep the toroids cool. I mounted the toroids on foam pads and stuck them down with silicone rubber. It is difficult to get suitable capacitors in the values required so the filter Cs are made up from smaller ones, the 12nFs are 10 + 2n2 and the 27nF is 22 + 4n7. At the filter output there should be a good sinewave at up to 900W into 50 ohms. Large deviations from 50 ohms will cause the filter components to get hot.

Choke L1 is made from a two-inch (50mm) piece of receiver type ferrite rod wound with about 20 turns of 1.5mm enamelled copper wire. The Hall-effect detector is

COMPONENTS

Table 1: Parts list for the main transmitter

Part	Type	Qty	Supplier	Part no	£ each part*
IC1	HEF4001	1	Farnell	384-434	0.60
IC2	HEF4017	1	Farnell	384-720	1.00
IC3	HEF4538	1	Farnell	384-630	0.70
IC4	HEF4013	1	Farnell	386-339	1.50
IC5	TC4426	1	Farnell	295-024	1.80
IC6	HEF4023	1	Farnell	384-847	0.50
IC7	7812	1	Farnell	413-215	0.58
TR1, 2, 3, 4	IRFP450 IR	4	Farnell	355-665	4.79
Alternative	IRFP450 Harris	4	Arrow	409356F	2.84
TR5	TIP115	1	Farnell	426-635	0.62
TR8	BF245A	1	Farnell	352-044	0.35
D1, D2	IN4936	2 (10)	Farnell	366-950	0.11
Hall effect	OHN3040U	1	Farnell	405-656	2.30
BR1, BR2	35A 600V	2	Farnell	234-151	3.17
LED	HE red	1	Farnell	637-087	0.15
Washers	TQ3P Heat pad	4	Farnell	936-753	0.56
RL1	12V coil	1	Arrow	429052G	3.92
C	10,000/100	2	Arrow	039366H	12.95
Clip for C	51mm	2	Arrow	013112C	0.66
C	22,000/25	1	Arrow	091958F	4.11
Clip for C	V3/H2	1	Arrow	013013R	0.52
C	1 μ /63	2	Farnell	228-710	0.06
C	100nF	10	Farnell	545-892	0.10
C	10 μ F/25	1	Farnell	920-502	0.03
C	22 μ F/25	1	Farnell	920-514	0.04
C	47 μ F/25	1	Farnell	580-466	0.10
C (swr br.)	22nF/250V	2	Farnell	303-8075	0.33
Rx filter C	2n2	2	Farnell	105-889	0.17
Rx filter C	56pF	1	Farnell	105-886	0.15
Rx filter C	20nF	1	Farnell	568-223	0.09
C	1nF	1	Farnell	635-147	0.07
Case	95x125x220	1	Farnell	518-876	11.0
Tx Filter Cs	10nF/1.6kV	2	Farnell	577-832	0.66
Tx Filter Cs	22nF/1.5kV	1	Farnell	106-371	0.35
Tx Filter Cs	2n2/2kV	3	Farnell	577-819	0.72
Tx Filter Cs	4n7/2kV	3	Farnell	577-820	0.59
Trim C	60pF	1	Farnell	303-872	2.64
C in parallel	47pF	1 (10)	Farnell	498-312	0.13
Decouple C	2 μ 2/250V	1	Farnell	148-896	1.11
Mains trans	2x35V 530W	1	Farnell	177-945	35.9
T1	58mm	1	Arrow	420605G	6.23
L1	Ferrite rod	1	Arrow	409653X	0.82
T2,T3	Toko	2	Bonex	719VXA-A017A0	0.72
T4, Ferrite	Tube 15x11	1	Maplin	AN78K	0.64
Filter cores	T200-2	2	Mainline	T200-2	4.30
Alternative	T200-2	2	Bonex	T200-2	
Meter	0.25mA FSD	2	Maplin	LB80B	3.40

Table 2: Parts list for the VFO

Part	Type	Qty	Supplier	Part no	£ each part*
Transistor	BC183	1	Farnell	356-580	0.15
choke	1mH	1	Farnell	608-609	0.25
C	22pF	1	Farnell	105-056	0.20
C	330pF	1	Farnell	105-063	0.20
C	1nF	6	Farnell	105-066	0.20

Unspecified parts, resistors and capacitors etc. are standard types.

*Subject to change

You will also need:

Lots of 0.4W resistors.
 VFO variable capacitor, about 100pF.
 Heat sink.
 Connecting wire and thick power wire.
 15A car fuse and holder.
 Tx switch.
 Shunt for 15A full scale.
 Key, aerial and RX sockets.
 A diode for the relay (1N4148 or similar)
 Silicone rubber sealer.
 Double sided copper-clad board for ground plane
 Two break-before-make 3 way rotary switches
 (one 3-pole).
 Short length of 2-core speaker cable.
 Some 1.5mm (16g) enam copper wire for 14V
 secondary.

Suppliers

Farnell 01132 636 311
 Arrow 01279 626 777
 Bonex 01753 549 502
 Mainline 01162 777 648

All the prices are exclusive of VAT and some of the cheaper items are subject to a minimum supply quantity so you may have to buy ten!

placed near one end of this rod and the spacing adjusted to trip at about 20 amps.

Remember: the FET source and drain connections and all others carrying the 100V line have to pass up to 20A and must be made using thick wire and good solid joints!

The power supply section of the rig is very simple with no regulation of the HT supply. Switch SW4 has been included to apply half volts to the PA for low power use and testing (it should be a heavy-duty switch or relay). The large smoothing capacitors are expensive but good ones are required to ensure a T9 note. The transformer as supplied does not have a 12V winding but this is easily added by winding 30 turns of 16SWG wire through the toroid. Both bridge rectifiers are the same although the 12V one could be a smaller one if desired. A simple 3 terminal regulator, IC7, provides plenty of capacity for the 12V relay and electronics. It can either be clamped to the case or to the ground plane copper clad board.

The fuse in the HT supply is one of the car type plug-in fuses that are available in 15A rating (blue).

Testing

Get the PSU, VFO and CMOS stages working first, you should be able to hear the 136kHz signal from the 4013 on your Rx by placing a probe near the chip. You can safely key the drive into the MOSFETs with no power applied to them (remove the fuse). Check with a scope that you have complementary 12V square waves on the two gates. Don't worry if the waveform is a little rounded off – it is due to the high gate capacitance of the FETs.

Connect the Tx to a 50Ω load and, having selected the first tap on SW2, apply 50V to the FETs (SW4 in low position) with a resistor in place of the fuse to limit the current. The FETs should draw no current when there is no drive. Press the key and the output stage should draw a few amps and produce a few watts into the dummy load. If the shut-down LED comes on you've either got a mismatch, the SWR bridge is connected backwards or it needs the 60pF capacitor adjusting. Any sparks or smoke at this stage is a bad sign.

If all seems well you can remove the current limiting resistor and continue tests, increasing the power by selecting taps. Key the rig in short bursts and check for overheating of FETs and cores – there should be none yet.

When you are happy that the Tx is working OK, load it up to 15A PA current and slowly move the Hall device nearer to the end of the ferrite rod (CH1) until the protection circuit trips. Move it just a tiny bit further away and fix in position with silicone rubber.

Never key the rig into open circuit, especially on the high HT voltage. Do not turn the tap switch under power, as it will spark and may damage the contacts – and possibly the FETs!

The Tx is sensitive to load and some method of measuring the impedance of the aerial system is desirable. The SWR bridge will give a useful indication of match but the power readings will need to be calibrated. Make sure you have the forward and reverse lines the right way round or the rig will shut down as soon as any power is produced! Trim the 60pF capacitor for minimum reflected reading when the rig is transmitting into a 50Ω dummy load. Always monitor the current to the output stage; try to keep the key-down current below 15A to be sure of staying within the limits of the FETs.

The receive filter will need tuning up. It is very sharp so use a signal near the centre of the band if possible.

Possible variations

Any reasonable high voltage power FETs should work; a low rds-on is desirable, as is a suitably large current rating and a voltage rating of four times the supply rail. I used IR ones in my Tx; I have not tried the cheaper Harris ones but others have used them. In my version I have two meters on the front, one permanently monitoring current and the other switchable between HT Volts, forward power and reverse power, but one would do if you can't find a source of cheap meters. I got mine at a rally for £2 each.

Any even multiple of output frequency could be used as a drive source. The 4013 must be left to do its divide by two but any other division ratio could be used ahead of it. It may also be OK to take in 136kHz and generate the complementary square waves with a squarer and inverter.

A 73kHz version works fine using exactly the same circuit. Just double all the component values in the output filter.

400W Power Amplifier using 572Bs

By Steve Rawlings GW4ALG

This power amplifier (PA) gives 400W output using a pair of 572B valves. The amplifier uses many of the components, including the cabinet and power supply, from an old KW1000 HF linear amplifier as shown in Fig 3.17.

Circuit Description

The circuit of this amplifier is shown in Fig 3.18. It was converted from a switch-tuned grounded-grid configuration to an aperiodic (untuned), push-pull amplifier, rather like the QRO Class B Modulators of the 1960s.

On receive, a voltage of about -25V, obtained from the 6.3VAC supply to the PA valve heaters via a voltage quadrupler, is used to bias the valves off.

Negative bias is used to reduce the quiescent standing current so that the two 572Bs now draw about 40mA (rather than the 100mA under 'zero bias' conditions). This is achieved with bias provided by low impedance negative voltage obtained by half wave rectification of the 6.3V AC supply to the PA valve heaters. These voltages are selected by contacts RL1a.

Construction

All the original tuning and switching components were removed and the holes in the front panel blanked over. The original 572Bs and their bases were retained but the original grounded grid arrangement was converted to a grounded cathode (heater) as shown in Fig 3.18. Toroidal transformers were constructed and used in both the grid and anode circuits, using cores of Philips 3C85 material (see Appendix).

The grid transformer (L1) is wound on a 25mm OD 3C85 ring core, with 8 turns on the primary, and 38 turns on the secondary (centre-tapped). With 2kV on the anodes, 10W of drive easily delivers 400W (at an anode current of 350mA).

The output (anode) transformer uses a 58mm OD 3C85 core with a primary winding of 140 turns, centre-tapped; and a secondary winding of 9 turns to provide a 50Ω secondary.

Following advice from Finbar EI0CF and Dave G3YMC I wound some layers of insulation on the core because of the high voltages that were to be expected on the primary.

I began with a few layers of PTFE tape around the toroid, which was then held in place by a few layers of white PVC tape. A further layer of PTFE tape and PVC tape were then added for good measure.

I then wound the centre-tapped anode winding with 140 turns of 26SWG enamelled copper, anchored with nylon cable ties, and a spacing of about 14mm between the start and finish points. More PVC tape was used to wrap the centre (but not the extreme ends) of the primary winding.

The secondary winding used wire from stripped-down mains flex. The turns were wound over the centre part of the winding, which had previously been covered with PVC tape. The transformer was then mounted on some 3mm Paxolin board.

The ends of the windings were sleeved with neoprene



sleeving and terminated on terminal posts fitted to the board.

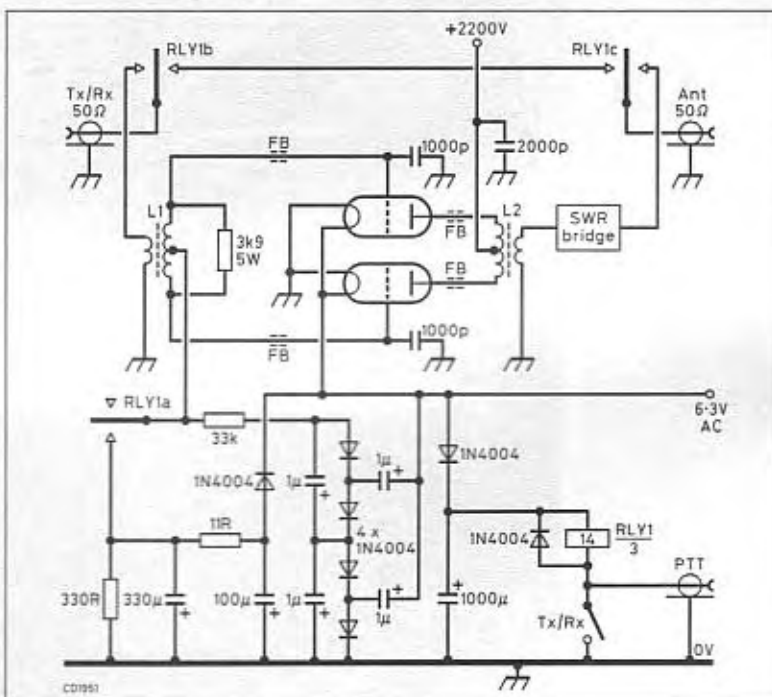
Initial tests

During dummy load tests the amplifier output was 400W in the frequency range 125-150kHz (the range of my transverter).

Initial tests with the amplifier were very exciting! Self-oscillation of the PA produced some colourful displays owing to high voltage flashover, both internal and external to the valve envelope! Purple sparks of 8mm length were observed within the valve and across the output transformer terminals. Eventually, sufficient padding of the grid circuit brought the PA to a stable condition (including 5 x ferrite beads on each grid lead, and a 1000pF capacitor from each grid to earth).

Fig 3.17: General view of the 572B LF transmitter using a converted KW1000 HF linear amplifier purchased for under £200. The original tuning and switching components have been removed and the holes in the front panel blanked over. The original SWR meter and switching has been retained.

Fig 3.18: The circuit diagram of the GW4ALG passive grid push-pull 572B amplifier.



LF Power Amplifiers using a Bridge Configuration

Peter Dodd G3LDO

As you will have noticed, most of the PA designs have evolved to use push-pull FETs in Class D.

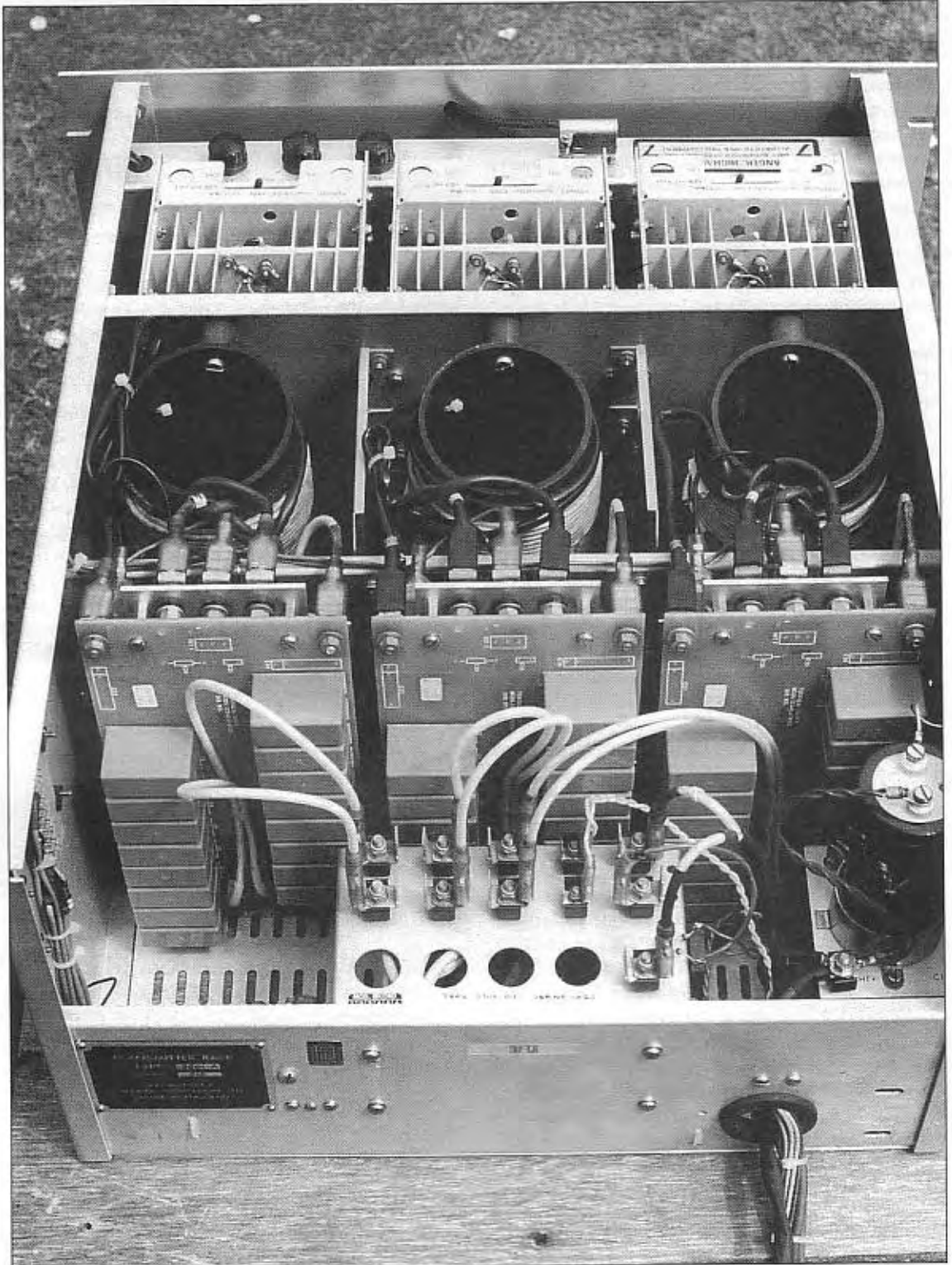
When the Decca stations were decommissioned and radio amateurs were given the opportunity of examining the Decca transmitters, it was found that a bridge configuration was used in the transmitter PA modules, rather than push-pull. Here the bridge configuration is

examined to see if anything can be learned from it.

A simplified diagram of the Decca transmitter is given in Fig 3.19. The driver circuit comprises a comparator IC1a operating as an inverted Schmitt-trigger, providing a squared output.

This squared output is fed to two parallel-connected inverters and then input to the gate of the IRF520 driver

Fig 3.20: The Decca transmitter chassis viewed from the rear. The inductors and capacitors that make up the tank circuits take up more room than the transmitter modules.



whose drain is connected to primary of the main driver transformer.

The power amplifier uses four IRF250s power FETs TR3 to TR6 connected in a bridge configuration. The gates of TR3 and TR5 are switched in opposite phase by the transformer-coupled drive from the driver board. Similarly, TR4 and TR6 are switched in opposite phase. The overall phasing is such that TR3 and TR6 conduct together during one half-cycle of drive and TR4 and TR5 conduct during the other half-cycle of drive.

Output is taken from centre-tapped and damped chokes L1 and L2.

The power FETs are protected by diodes. In addition the bridge circuit includes two voltage dependent resistors for transient suppression.

Typical ON resistance for IRF250s is 0.085 ohms at 25°C and 0.17 ohms at 150°C.

Output from the power amplifier module is to a tank circuit. The tank circuit, which is series-tuned, converts the square wave output from the power amplifier to a sinewave.

The Decca transmitter chassis contains three amplifier modules described above; see Fig 3.20. The outputs from the three tank circuits are combined by the output transformer, whose secondary provides a 71Ω output.

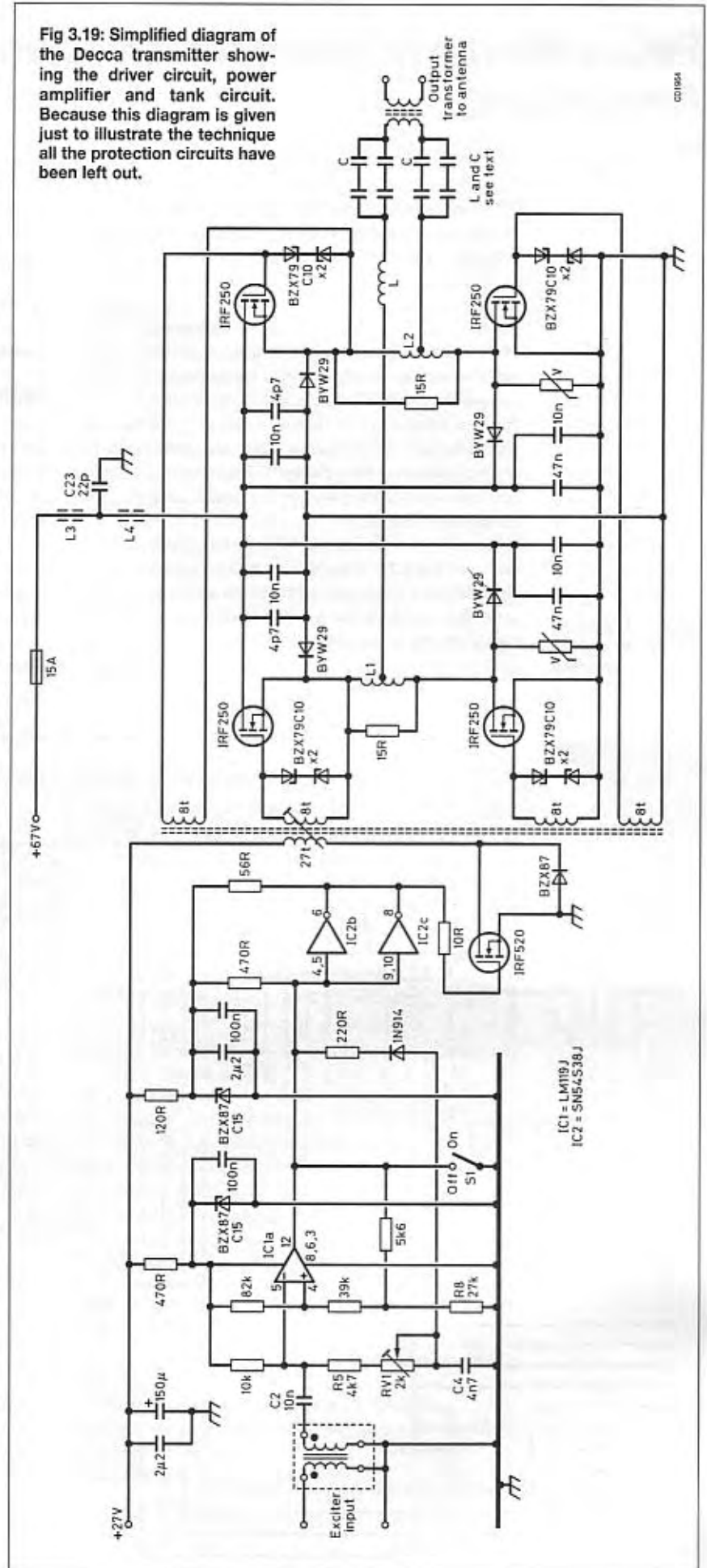
These transmitters are designed to operate continuously using only remote supervision. The modules are closely packed together with no forced air cooling and very little conduction cooling so they must be reasonably efficient and, above all, reliable.

So is there any advantage to the bridge configuration? Well apparently its main benefit is the same as in bridged audio amplifiers: four times the power output for a given supply voltage and load impedance. If the power is taken out via an RF transformer, *each* end of the primary winding alternates between full supply rail and ground. This is the same as applying twice the voltage to a winding with one end grounded, without actually having to stress the amplifying devices with twice the voltage. To achieve the same power in a non-bridged arrangement, one must go to a lower load impedance and deal with higher currents; or change the turns ratio and operate the switching devices at higher voltage, or some other compromise.

The design would appear to be able to deliver the required power output with a mere 67 volts supply. This implies a safety consideration that most amateurs ignore.

One possible disadvantage is the relative complexity of the driver transformer. This could be circumvented by using a HIP4080 high frequency full bridge FET driver as shown in the Appendix. A further disadvantage might be the size and complexity of the tank circuit as shown in Fig 3.20.

Fig 3.19: Simplified diagram of the Decca transmitter showing the driver circuit, power amplifier and tank circuit. Because this diagram is given just to illustrate the technique all the protection circuits have been left out.



Pulse Width Modulator for Class D Power Amplifiers

By A C Talbot G4JNT

The widespread use of high efficiency Class D switching power amplifiers for transmitting CW on LF has meant the near impossibility of employing these for many of the other efficient digital communication modes around now. These modes, such as PSK31/PSK08, all require a linear amplifier to preserve the amplitude shape of the waveform. One way of overcoming this problem whilst retaining switching mode transmitters is to separate out the amplitude envelope and phase components of the drive signal and treat these separately. The phase component is applied to a conventional phase modulator in the drive signal to the Power Amplifier, whilst the amplitude envelope is applied to a high power modulator controlling the supply to the PA.

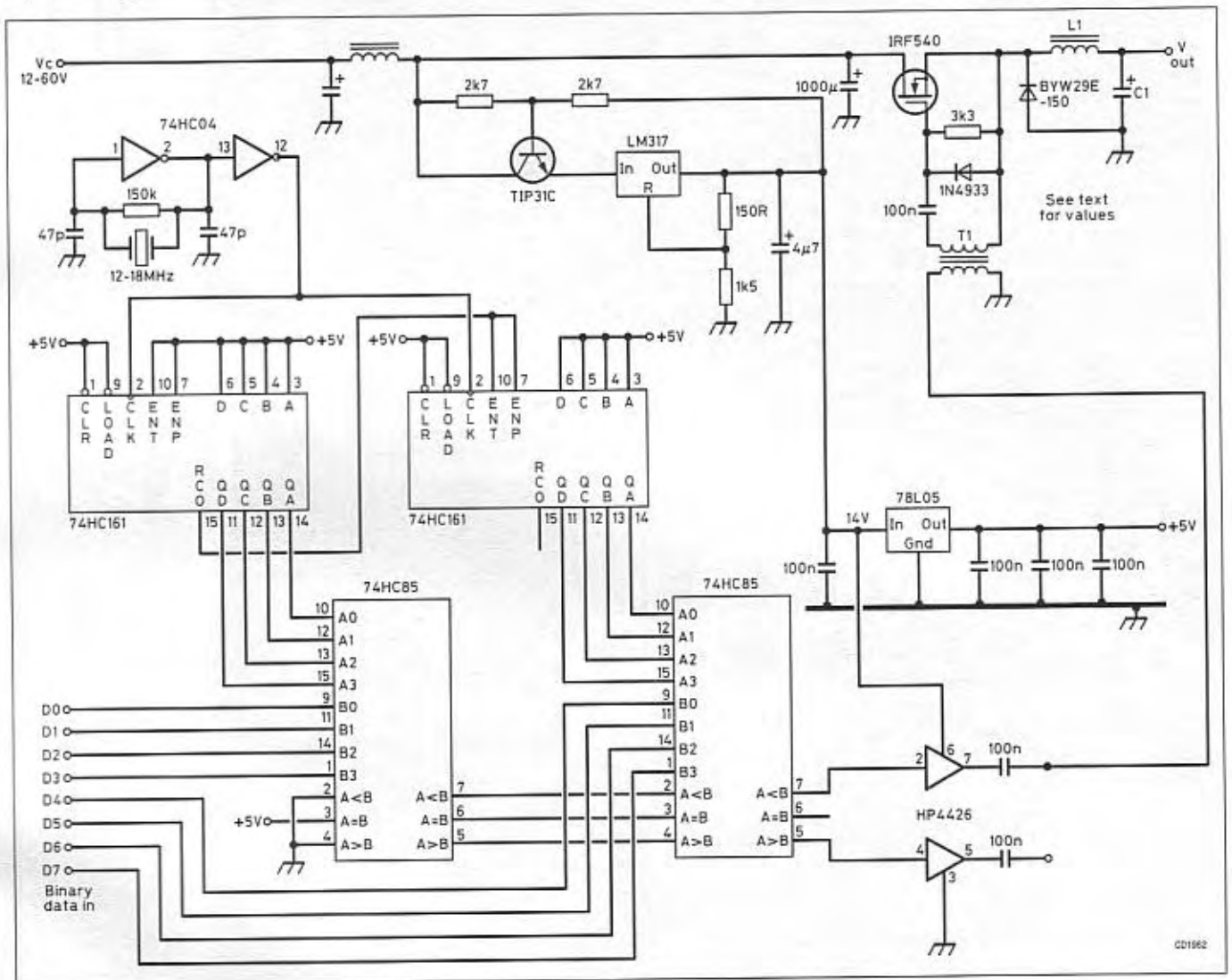
This design is for a Pulse Width Modulator (PWM) is shown in Fig 3.21. It has been tested for supply voltages up to 50 volts at currents of up to 10 amps, but there is sufficient margin in the specified components for even higher current. It is equally suitable for lower supply voltages, and this makes possible PSK31 or similar beacons

by generating separate phase and amplitude signals from a custom keyer module. PW Modulation is a very efficient means of producing a controllable DC level at high currents as no power is wasted in resistive droppers. The DC input waveform is chopped and converted back to a clean waveform by an inductor and capacitor; PWM forms the basis of all Switch Mode Power Supplies.

PWM Generation

This design can be separated into two functions. The first part generates a variable pulse width switching signal at a frequency of around 50kHz. The pulse width is controlled by an eight-bit digital word. Although analogue PWM circuitry was considered, because the beacon application for this was originally intended would provide a digital word for amplitude, a simpler solution would be an all-digital PWM generator. If an analogue input is needed, a simple one-chip A/D converter could be added to make the conversion. A final switching frequency of 50kHz was adopted; although a higher value would be preferable it is

Fig 3.21: The G4JNT pulse width modulator, circuit diagram.



obviously desirable to keep away from 70kHz and sub multiples of 137kHz. 100kHz, where many modern SMPSUs operate, was a possibility but would require fast logic circuits for 8-bit resolution and so it was decided to stay with 50kHz. The exact frequency is not critical and any value in the range 30-80kHz will work, but see the notes on output inductor sizing later.

The PWM function is generated very simply by driving a synchronous 8-bit counter with 256 times the final switching frequency – in this case 12.8MHz – so the final output is at 50kHz. This signal is then compared at any instant with the command word in an 8-bit comparator. The comparator's A < B output (or A > B, depending on polarity) is then the required PWM output signal. This is applied to an HP4426 chip for driving the switching MOSFET. The circuit could be further simplified by using a single chip 74HC682 8-bit magnitude comparator rather than two 4-bit devices, but at around £4 for the 8-bit device compared with 2 x 34p for the 4-bit ones the reason is obvious! Further simplification would be possible by using a packaged crystal oscillator module, but the discrete design means any surplus crystal can be pressed into service.

A voltage regulator is included to provide a 14V supply for the HP4426 driver chips from the 50V input line. A two-stage regulator is needed as 50V exceeds the input limit for a single LM317. The TIP31 power transistor shares the voltage drop equally between itself and the voltage regulator IC, and means that input voltages up to 70 to 80 volts can be catered for. The driver takes around 25mA, and as the two regulator devices dissipate around half a watt each, a small heatsink is necessary.

Power Switcher

The switching circuitry is very simple in concept, but some careful component selection is needed to ensure reliability and correct operation. A conventional N-Channel MOSFET is used with a pulse transformer and DC restoration circuitry to perform the level shifting. The pulse transformer driving the FET is a simple 1:1 device made by winding approximately 40 + 40 turns on a suitable ferrite core, eg RS 212-0869. A P-Channel device would remove the need for the transformer but in turn would need a custom drive circuit rather than the simple chip used here. Furthermore, P-Channel devices have higher 'on' resistances for a given rating and cost more than their N-Channel equivalents. The IRF540 specified is rated at 100 volts which is satisfactory for inputs up to 50V; for an input below 25V, 50V devices are adequate. R_{ds(on)} for the IRF540 is quoted as 0.08 ohms giving minimal losses at 10A. The commutation diode D1, whose purpose is to maintain current flow during the switch-off period, is not critical provided it is a fast switching device capable of carrying the required output current. These devices are very plentiful in SMPSUs and are low cost; the device specified is overkill but I had a lot of these surplus.

By far the most critical component is the output inductor and these design guidelines must be adhered to for proper safe operation. It causes the largest loss in the modulator, and a good test of proper functioning is to check its temperature rise after a long period of operation. Its job is to maintain the output current flowing during off periods (via the commutation diode), and needs to contain enough stored energy during this period to allow this. In addition, since there is a large DC component flowing

through it the inductor must not saturate due to this or the volt-second product during the 'on' period. The DC term dictates that iron dust toroids or gapped ferrite cores are used. Gapped ferrite cores suitable for SMPSUs are not so available from the popular component suppliers as was once the case, but fortunately, a range of suitable iron dust toroid cores for just this purpose are available. The inductance value needed is based on the *minimum* output current wanted at the regulated voltage. If, during the 'off' period, the voltage across the inductor falls to zero regulation suffers and at the minimum value output current rises. For a resistive load – such as a class D switching PA – this is not a real problem as here the waveform will be close to its minimum amplitude for a short period only.

The minimum inductor value has the highest value at 50% duty cycle, where V_{out} is half the supply voltage, and L_{min} is given by

$$L_{min} = 0.25 \cdot V_{in} / (F \cdot I_{min})$$

Defining the minimum load current I_{min} without knowing in advance what signals will be transmitted is difficult, but as a rule of thumb to start with this can be taken to be 10 percent of I_{max} and adjustments made later after testing. For a 50V, 10A supply, I_{min} becomes 1A and L_{min} = 0.25 x 50 / (50000 x 1) = 250µH. RS supply a range of powdered iron toroids and the largest one of these (reference 232-9404) is a 33mm diameter device with specific inductance A1 value of 60nH/turn². To reduce the number of turns of thick wire needed two cores are stacked, which doubles the A1 value. The number of turns is then given by SQRT (250µH / 60nH) = 64 turns. The wire used needs to be capable of carrying the full output current and I used two parallel conductors of 1mm diameter. Since there is a large high frequency component as well, the use of Litz wire or multiple strands of thinner conductor will be better. This number of turns near enough fills the cores but the resistance is satisfactory. When making the prototype I ran out of wire after 40 turns so the value ended up at 192µH, but for breadboarding this was adequate. A range of ready-made SMPSU inductors are available from RS and these should prove adequate although none have been tested.

The final component of note is the output capacitor C1. This acts in an identical manner to the capacitor in a normal power supply, and defines the ripple voltage depending on load current.

Its value is given by $C = I_{max} / (Freq) \cdot V_{ripple}$.

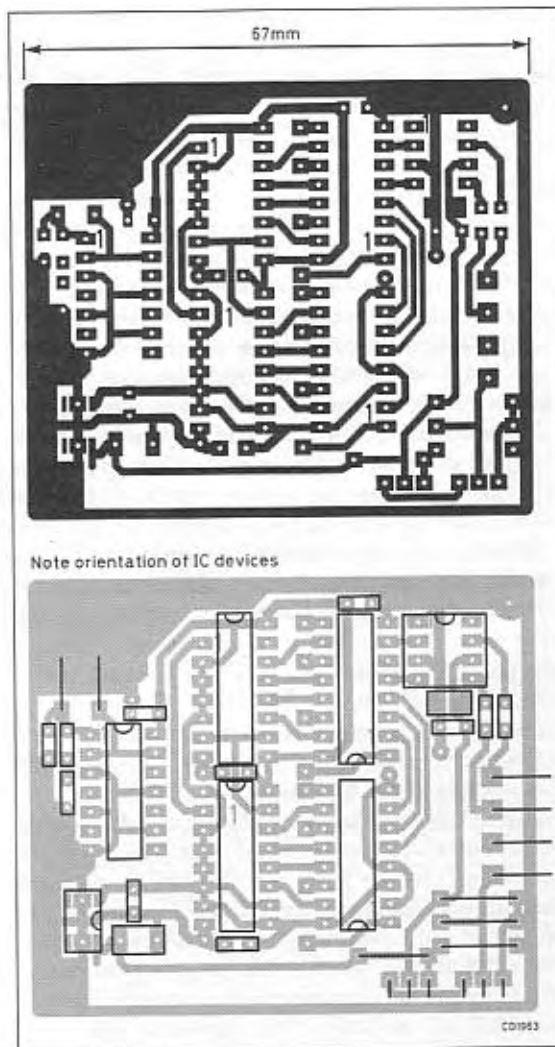
If the ripple is arbitrarily set at 0.2V, then for 10A load C = 1000µ. A low series resistance device suitable for SMPSU design is needed and obviously of the required voltage rating. RS type 191-7987 rated at 63V is just one of a large number suitable for the purpose.

Results

The complete modulator using the components described was tested using a six bit digitally generated PSK31 envelope (the two LSB inputs were tied to 0V) from a 50V rail driving a pair of 10Ω 250 watt fan-cooled resistors. After two hours operation:

Temperature of IRF540 and BYV29-400 mounted on a 100mm square heatsink	45 deg (warm)
Temperature of inductor core	55 deg (warm to just hot)
Temperature of load	Ouch!
Overall efficiency when using a typical PSK31 beacon	

Fig 3.22: The G4JNT pulse width modulator, component and PCB layout. Note the orientation of the HC85 and HC161 chips on the PCB.



message waveform looked to be of the order of 85% to 90%. Owing to the lower-than-intended value of inductor, the reconstituted voltage waveform across the load did not quite reach its minimum possible, but from a peak of 48V it did get down to 5V so in practice this should be quite adequate. For waveforms with a higher rate of change than PSK31 the output capacitor may need to be smaller, increasing the ripple component, and a trade-off will be needed here. Too much ripple could get into the PA stage and appear as spurious outputs, leading to an interesting concept. By making the switching frequency equal to the RF carrier, no spurious products will appear. While 137kHz is a bit on the high side for the design as it stands, it could be made to work with the right choice of ferrite materials. At 73kHz no changes are needed.

For lower output voltages and currents the inductor, and possibly output capacitor, will need to be re-calculated as shown above. As I have not yet built a class D transmitter, no full system tests have yet been tried, but the PSK31 beacon was tested on Top Band using a classic (and very inefficient!) emitter-follower modulator and yielded a signal almost indistinguishable from a conventionally generated waveform. A lower input voltage will also allow simplification of the voltage regulator elements. For supplies less than 30V, the TIP31C can be dispensed with along with the two 2k7 voltage-sharing resistors. For operation below 15V the LM317 may be dispensed with.

Note that it is not possible to turn on the device fully to achieve absolute maximum output with a continuous

binary drive of 255 or '11111111' as the resulting DC drive level cannot pass through the pulse transformer. A value of 254 or '11111110' works, however, as the small negative pulse is enough to ensure proper functioning of the DC restorer on the gate of the switching MOSFET.

Construction

A printed circuit board layout for the PWM and voltage regulator components is shown in Fig 3.22. This employs 'semi' surface mount construction where traditional components are mounted on top of the board, ICs having their legs cut flush or carefully bent underneath to allow this. The layout of pads allows use of either 1206 SMT components or smaller wire ended devices. The underside consists of a ground plane and connections through are made by drilled holes where shown.

Note the orientation of the HC85 and HC161 chips on the PCB. Optimum layout dictated that these should sit in opposite directions – that's discrete logic for you!

The power switching components, including the pulse transformer, are not included on the PCB as the layout is dependant on the final wanted configuration of voltage and current. Layout should follow good high frequency practice with short fat leads carrying the high current switching signals.

Polarity of the windings on T1 determine the drive sense. Reversing either winding (or using the alternative comparator output on the logic board), will result in swapping over the drive sense so that '254' will give minimum output and '0000001' maximum. The easiest way to make sure that polarity is correct is simply to try it!

4

Low Frequency Antennas

LF Antennas

By Bob Vernal ZL2CA

Most of the text books on commercial LF transmitting antennas mention very high towers and high powered transmitters, and place considerable emphasis on ground systems. However, the amateur constructor is hardly in a position to erect huge verticals or to run a large number of long radials, so while the relatively efficient LF transmitting scenarios found in text books are interesting they are generally well beyond the capability of an amateur station.

While radiation efficiency of an amateur LF antenna will always be well under 1%, it is worthwhile considering a range of factors in optimising an antenna, as improvements to antennas are a key to success with LF experimentation.

Polarisation

Vertical polarisation is basically the only practical way of transmitting on the lower bands, as horizontal polarisation, if tried, would be largely 'shorted out' by the conductivity of the ground. Basic choices for amateur construction of a vertically polarised LF antenna, in summary, are:

- A loop (magnetic) antenna, in a vertical plane. This has a directional figure eight pattern with a null off each side of the loop. Needs to be series resonated with suitable capacitance (with high voltage and current rating), range of tuning capacitance up to about 10,000 picofarads. Input resistance, when tuned, is typically in the range 1 to 10 ohms (mainly depending on wire gauge, tuning loss and ground coupling). Radiation resistance (and hence efficiency) depends mainly on loop area, and looks a lot better when units are given in milliohms.
- An insulated vertical (electric) antenna fed against ground. Omnidirectional pattern (equally poor in all directions?). Radials are desirable. Some top loading is highly desirable. Needs to be series resonated with a suitable inductor, range of tuning inductance up to a few millihenries. Input resistance, when tuned, is typically in the range 10 to 100 ohms (mainly depending on ground resistance and tuning loss). Radiation resistance (and hence efficiency) depends mainly on effective height, and like LF loops, looks a lot better when units are given in milliohms.

Both loops and verticals have been used by amateurs; however, the vertical has proved to be the most practical and efficient.

The LF Vertical Antenna

As a first approximation, an electrically small LF vertical antenna is equivalent to a low value capacitance of typically a few hundred picofarads. At LF the capacitive reactance is rather high, of the order of thousands of ohms. Useful transmission needs significant current to flow upwards from the base of the vertical, so the high inherent capacitive reactance needs to be series resonated using a low loss (high Q) loading coil. The RF current that flows at resonance then depends on the value of remaining resistance. The system loaded Q can be quite high, so tuning can be critical to a kilohertz or so.

A short vertical has a current distribution that tapers fairly linearly from being maximum at the base to zero at the top (the current distribution eventually becomes sinusoidal for electrically longer antennas). The DX signal strength directly depends on the product of antenna current and effective height of the antenna (the more amp metres, the better). The linear tapering of current in the short vertical means that the effective height is theoretically half of the physical height.

Practical tests have consistently confirmed that an electrically short pure vertical can be considerably improved in efficiency by introducing some 'top loading'. The T type of antenna is a classic way of doing this, with an inverted L being second choice. The T antenna tends to provide net cancellation of horizontal polarisation components, as the current going up the vertical splits fairly evenly and flows in opposite directions towards the end of the wires in the top loading, whereas the inverted L involves some net horizontal component from current in the top wire.

A T type of antenna is suited to being suspended between two supports, with ideally the vertical drop wire being taken from the centre of the top span. However, variants on the T or inverted L are also practical, and implementation of an amateur constructed LF antenna is basically a case of making the most out of existing masts and taller trees to support a useful amount of top loading.

A T type of antenna can be considered as having a certain capacitance, say C1, to ground from the vertical part, and another capacitance, say C2, to ground from the top loading part. Ideally the capacitance C2 of the top part should be many times the capacitance C1 of the vertical part. The current will still taper off from being a maximum at the bottom feedpoint to being zero at the end of each top loading wire, but the higher the ratio of C2/C1

then the less tapering off in current will occur in the vertical part (and it is mainly the amp metres in the vertical part which generates the DX signal strength). Thus the effective height of a T antenna is more than that of a vertical of the same physical height. The theoretical limit is for the effective height of a T antenna to approach the physical height (when approaching infinite top loading).

Adding capacitive top loading to an electrically short vertical antenna actually has multiple advantages, namely:

- it significantly increases the effective height (by up to two times).
- the higher resulting system capacitance (approaching $C_1 + C_2$) spreads the electric flux and reduces the density of some ground currents.
- the radiation resistance increases (but is inter-related to the previous two points, where different concepts lead to similar conclusions).
- tuning inductance is less, so there will be less tuning coil loss.
- a lower L/C ratio gives a lower loaded Q than for a pure vertical, and obtaining lower Q allows for wider operating bandwidth, which can otherwise be a problem at LF.
- the RF voltage applied to the antenna is also related to the loaded Q, so a lower Q means that insulation is not so critical, or alternatively, more transmitter power can be applied for a given breakdown limit.

Adding inductive loading within the antenna itself is not so attractive as adding capacitive loading. There is very little point in putting an inductor near the top end of a vertical, as there is little current flowing at that part of the antenna. In theory, centre loading of a pure vertical gives some 3dB advantage over base loading. With a T antenna, tests have shown that placing a loading coil between the top of the vertical wire and where the top loading connects can improve performance by about 1dB. However, it is easier said than done to physically support a low loss coil of the order of 1 millihenry and adequately weatherproof it for fairly high RF voltage. A practical and widely used approach is to place the loading coil near ground level, in a weatherproof enclosure, as this allows the use of a relatively large coil, possibly with integral variometer, as well as allowing for optimum matching to a given feeder impedance (say 50Ω coaxial cable).

Capacitance of Wires Above Ground

A lot of detailed information has been published on the capacitance of various configurations of wires above ground. Without getting into too much detail, the key physical factor is that there is a logarithmic variation of capacitance with both diameter of wire and height of wire above ground level. Results of capacitance calculations based on reference values of 1 mm diameter wire and a T antenna with a height of 10 metres above ground are:

- vertical uplead approximately 6 picofarads per metre of wire.
- top loading approximately 5 picofarads per metre of wire.

A calculation for doubling the wire diameter from 1 to 2mm results in an increase in capacitance of some 7%. Lowering the height by a factor of 2, from 10 to 5 metres above ground, similarly results in an increase in capacitance of some 7%. This logarithmic relationship has the practical implication that neither the wire diameter used nor the height above ground level are critical factors in

determining the net capacitance. The length of the top loading wire is the main factor in determining net capacitance.

Further calculations were carried out to explore the impact of running two same diameter parallel wires, rather than a single wire. The reference condition again being wire of 1mm diameter at 10 metres above ground. The following results were obtained:

- 2 wires separated by 1 mm increases capacitance by about 4%.
- 2 wires separated by 10 mm increases capacitance by about 19%.
- 2 wires separated by 100 mm increases capacitance by about 39%.
- 2 wires separated by 1 metre increases capacitance by about 68%.
- 2 widely separated wires increases capacitance by up to 100% *ie* double the capacitance of a single wire, as one would expect.
- as found earlier, 1 wire increased in diameter from 1 to 2 mm, increases capacitance by about 7%.

The results indicate how proximity effect limits the realisable net capacitance for closer spaced wires. However, despite proximity effect, it is generally the case that "two wires are better than one" when it comes to increasing the net capacitance. Note that doubling the wire diameter results in four times the surface area, and happens to result in 7% more capacitance, whereas two same diameter wires that are nearly touching involve two times the surface area and happen to have a 4% increase in capacitance. Note also that doubling the wire diameter gives a weight increase of four times, so in terms of mechanical support and sag of wires in an antenna, two separate wires are generally better than a single thicker wire. It is clear that multiple wires are the answer to arranging useful top loading. It is also clear that this is already known to the designers of LF beacon antennas, as T antennas are extensively used, with typically two or three parallel wires in the top loading, with spacing of a metre or so.

There are proximity effects around the centre of the T where the vertical joins the top loading, so the net capacitance is actually moderately less than the sum of the capacitance of each wire in isolation.

Multiple Grounding

Multiple grounding is a topic that is not mentioned very often in amateur radio literature. A similar situation to LF grounding arises with grounding for mains supplies and for lightning protection. Consider an example of a single driven earth rod, in damp earth, having a resistance to ground of 30 ohms. If this were the sole ground connection for a vertical antenna fed with one amp, then the power lost in the ground would be $P = I^2R = 1 \times 1 \times 30 = 30$ watts. Now consider adding two more driven earth rods, spaced sufficiently so as to have independent current paths, and assume each then has the same RF resistance to ground as the first rod, *ie* 30 ohms. If the rods are treated as a single earth with three parallel connected rods, the combined resistance to earth is $30 / 3 = 10$ ohms. Again, feeding the antenna with one amp, the three parallel connected earth rods result in power lost in the ground of 10 watts, which is a third of that found for one rod alone. Thus compared to a single rod, the total power loss in the ground goes down by the same factor as the number of parallel connected rods. Another way of looking at this

example is to consider the current per individual rod, which is now 0.33 amp, and the ground resistance is still 30 ohms, so the power loss per rod is $0.332 \times 30 = 3.33$ watts. For all three rods, the total power lost is 10 watts, which of course agrees with the result already found.

The above example has used simple numbers to illustrate the basic theory of multiple grounding; however in practice there are unavoidable proximity affects between driven rods at practical spacing, which distorts the ideal radial pattern of current flow, with the consequence that the net resistance is not quite as low as simple parallel calculation indicates. This still results in the net resistance to ground of multiple connected driven rods being significantly less than the resistance of a single driven rod. Thus multiple grounding appears to be a significant technique for reducing ground loss in amateur LF antenna systems.

Another way of considering this reduction in ground losses with multiple connected rods is that it comes about from reduced current density, by spreading the current into several patches rather than one. Having multiple buried bare wire radials has a similar result in reducing current density by firstly dividing current between radials and secondly by dispersing these divided currents over the length of the radial. If multiple driven earth rods are used, the most useful place to use rods could be at the end of radials rather than at the antenna base or some other central part of the earth system.

Practical Results of the T Antenna used by ZL2CA

Several measurements have been done on a variant of a T antenna. The property is slightly sloping, and happens to have a house that clogs up some of the clearance below the antenna, so heights are somewhat relative. The vertical feed is fairly central and rises about 8 metres to where all top loading is connected to a fibreglass standoff from an earthed pipe mast. Three other supports are around 9 to 10 metres high. Approximately 120 metres of wire make up the top loading. A home-made RF impedance bridge indicates readings at 181kHz of $11 - j1200$ ohms (11 ohms and 800pF). The relatively low value of resistance is due to some combination of having many earth radials and distributed driven earth rods. The antenna capacitance includes some undesirable but unavoidable shunting by the earthed mast and proximity of roofing iron.

Loading coil RF resistance is about 4 ohms, so the resistance of the tuned antenna system is some $11 + 4 = 15$ ohms. When the transmitter is connected, this is equivalent to a further series resistance of about 5 ohms, so the transmitting setup has a total series resistance of some 20 ohms. A loaded Q test by tuning across to find 3dB current reduction bandwidth indicated a loaded Q value of just over 60, which checks out well for 20 ohms net resistance and 1200 ohms reactance. Also the antenna current is some 2.5 amps as verified on an RF ammeter for nominal 100 watts applied power, and this again checks out reasonably as 2.5 amps into 15 ohms involves a power of 94 watts. The transmitter is designed for 50 ohms loading, so is not optimally loaded when fed in series with the loading coil (15 ohms). Maximum antenna current was found to occur by feeding the transmitter at a coil tap that is 2 turns from the cold end of the loading coil (antenna at 57 turns). More recently the same result has been obtained by series feeding using a 50 to 15Ω low-pass L network, rather than tapping the coil. The RF voltage

applied to the vertical is some 3kV RMS for 100 watts transmitter power. Clearly there are safety precautions to observe.

The radiation resistance is unknown, but it would not be more than about 0.03 ohms (30 milliohms), even with a tail wind. 0.03 ohms as a fraction of 20 ohms net system resistance gives a radiation efficiency of no more than 0.15%. So 100 watts applied gives no more than 150mW radiated. QRP radiation maybe, but this gets consistent signal reports from Christchurch and Dunedin (each mostly sea path) and by skywave (night time, and QRN permitting) to many parts of New Zealand. The [NZ] regulatory limit of 5 watts EIRP is not realistically under threat, even with a higher-powered transmitter under construction.

In summary, an optimised LF T antenna should consider the following:

- have only a single vertical wire, or if multiple vertical wires cannot be avoided then they should taper towards the bottom feedpoint.
- have multiple wires in the top loading, the more wires the better.
- height and extent of top loading depend strongly on the site available, and supporting poles or trees (the higher the better for the vertical and the wider the better for top loading).
- top loading is likely best if kept entirely elevated, and drooping wires or inverted Vee type of loading may invite more losses than gains.
- try to minimise stray capacitance from the feedpoint and vertical wire(s) to ground ie the loading coil should be directly below the vertical part of the T.
- use as many ground radials and separate earth rods as are practical.
- there could be RF absorption from foliage, so keep reasonably clear of the greenery.
- use adequate insulation for all parts of the antenna, and observe safe clearance when transmitting.

Efficiency Improvement Measurements

An experimental way of investigating if antenna efficiency is better or worse is to carry out before and after measurements of antenna feed point impedance for a given trial condition. The loss resistance of an amateur LF antenna system is many ohms (say in the range of 10 to 50 ohms) and swamps the radiation resistance (several milliohms). The bulk of the loss resistance is due to ground loss. A decrease in the resistive component of feed point impedance means an increase in the radiation efficiency (and similarly an increase in resistance means lower efficiency). Thus careful measurement of feed point impedance, and with attention to the resistive component, is a way of assessing whether or not changes do improve radiation efficiency. So I tried this out.

My home made impedance bridge has a 100Ω linear pot as the variable series resistive element. I already knew that the antenna impedance was about 11 ohms resistance in series with 760pF. The marked scale for the 100Ω pot is inadequate for trying to look for small changes around 10 ohms. For the experiment I added a 15Ω resistor in parallel with the 100Ω pot, so that the combination provided a sort of expanded scale use of the 100 ohms scale for checking small changes around 10 ohms. What counted was not the actual accuracy of the impedance bridge, but rather the ability to carry out difference readings before and after adding top loading. I used an accurate

digital multimeter to find the actual value of the 15 Ω shunt, and also to verify that the 100 ohms pot scale was close enough to direct reading. The resistance at LF is close to those found from DC multimeter measurement.

For a test frequency of 182kHz, the "before" impedance was checked as 10.85 ohms in series with 760pF. I then added three more top loading wires, fitting them between existing top loading wires, which was easy to do for a trial. The "after" impedance was checked as 9.98 ohms in series with 805pF. Thus adding the loading resulted in a capacitance increase of 45pF (+6%) and a resistance drop of about 0.9 ohms (-9%). On retuning the loading coil and applying the usual 100 watts of power I also found that antenna current was in the range of 5-10%

more than for the initial top loading, which was another way of showing that the system resistance had been made a little lower (but includes coil loss and transmitter source resistance as well). Thus the experiment provided reasonable evidence that overall losses reduced by extending top loading.

While the impedance test set-up was on the bench, I also added a driven rod at the far ends of each of two radials wires, but I did not notice a drop in the value of the resistive component. My buried radials and multiple earthing was already known to be working fairly well, so the moderate addition of two rods was perhaps in the area of diminishing returns.

A Guide to LF Antennas

By Peter Dodd G3LDO

The Professional Approach

As noted by ZL2CA, earlier in this chapter, the insulated vertical (electric) antenna fed against ground has proved to be the most practical and efficient antenna on LF. The traditional LF antenna is normally a T, as shown in Fig 4.1, or an inverted L; these types of structures have come to us from commercial designs and are based on sound radio engineering principles. They have an omni-directional polar diagram as shown in Fig 4.2, with most of the energy radiated at a low angle, see Fig 4.3.

Amateur Limitations

For many radio amateurs the type of structure shown in Fig 4.1 is out of the question and certainly was the case at my location. The problem with high capacity top antennas is that they require a fair degree of antenna engineering to put them up (and keep them up). They also have a fairly high visual impact. Under the circumstances many of us were forced to look at other antenna configurations.

In the early days of 73kHz experimenting G4GVC put up a 100m long single wire antenna that was only 3metres high at the feed end 12metres high at the far end. This antenna worked far better than expected and computer analysis – see Fig 4.4 – showed that its energy distribution is not that different from the ideal shown in Fig 4.1.

There is a ground wave and a skywave component to radio signals, and on HF the skywave is predominant. On LF the groundwave is more predominant at the distances normally used. The characteristics of the conventional LF antenna are well suited to the LF commercial operator whose station performance would be degraded by any skywave component in the signal. On a broadcast station it causes distortion of the signal and on a navigation aid it can introduce errors.

For the amateur operator the requirements are very different. The objective is to communicate the furthest possible distance and this may involve skywaves.

Indeed, one of the most successful antennas is that used by OH1TN. This has a 28m vertical section with single wire 500m top, dropping down to 20m. As you can see in Fig 4.5, although the antenna is broadly omni-directional, a sizeable amount of energy is high-angle radiation;

anathema to the professional LF radio engineer.

Walter Blanchard G3JKV gave an explanation of why our horizontal wire antennas might perform better than we were originally led to believe. "...even if your horizontal antenna is only 10 or 12m high and slopes it still has that 10 or 12m as a vertical component and the rest is capacitive loading. LF penetrates the earth quite well (*cf* mine communications and even in salt water you can receive LF signals down to 10 or 12m) so if you have no

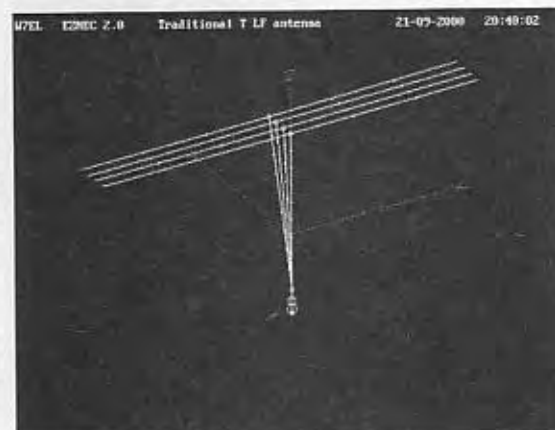


Fig 4.1: The LF T antenna.

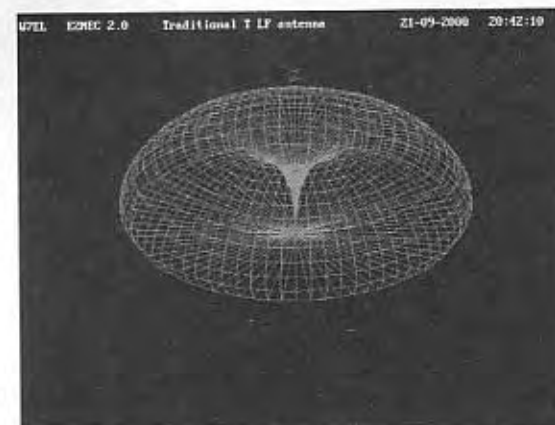


Fig 4.2: Three dimensional view of the energy distribution from a traditional LF T antenna.

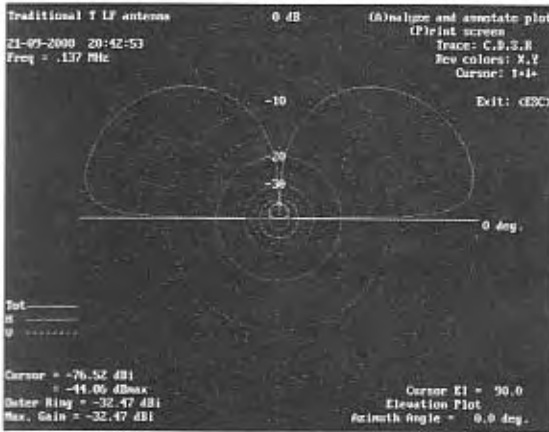


Fig 4.3: Cross-section of the energy distribution shown in Fig 4.2. This elevation projection allows easy analysis of antenna performance and shows the greatest gain at an angle of 13 degrees. The radiation is mainly vertically polarized with just a tiny component of horizontal polarization shown by the circle in the middle of the diagram (48dB down on the distribution at 13 degrees).

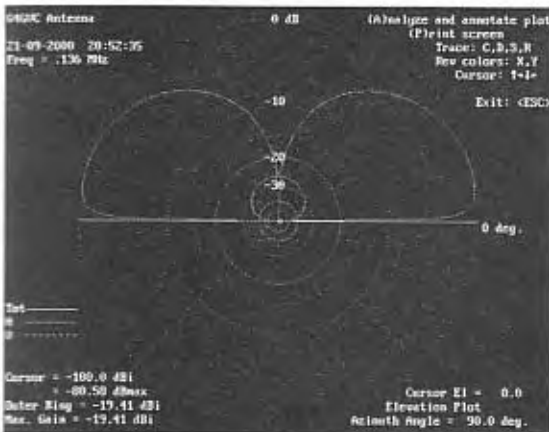


Fig 4.4: Energy distribution of the original G4GVC antenna. This shows a small horizontal polarization component 18dB down on the main lobe.

earth system your *virtual antenna* is probably buried some 18 or 20m, making the vertical bit perhaps 30m. If you then put in a copper mat on the surface you lose this and come back to the height of your wire, thus a weaker signal. So why do professionals put in their very expensive earth systems? Because the *virtual antenna* is totally unpredictable, varying with season, rainfall, water table height, earth conductivity, and so on.



Fig 4.5: Energy distribution of the LF antenna at OH1N. This antenna has a larger horizontal polarisation component than the other antennas.

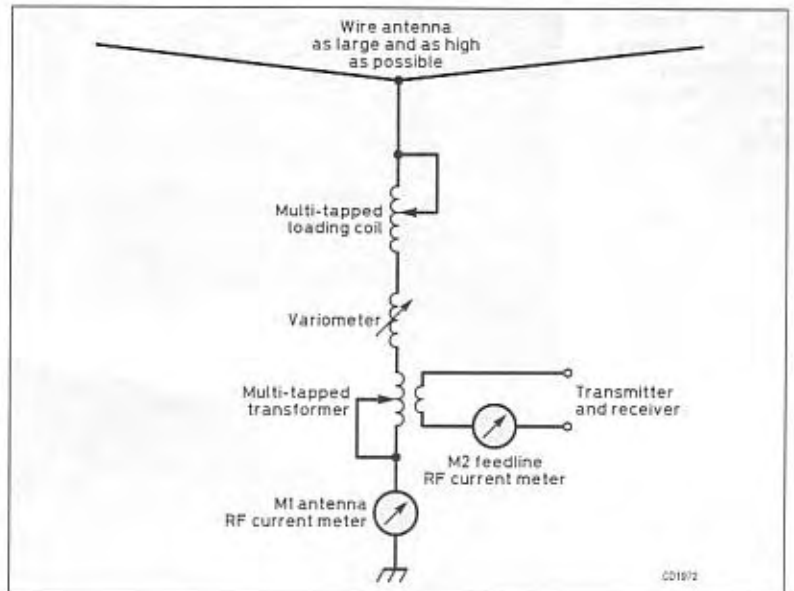


Fig 4.6: A typical LF antenna system.

Amateurs have the terrific advantage, they don't have to provide a 24-hour guaranteed service! A horizontal antenna is ideal for this situation – it puts out large amounts of skywave even if somewhat unpredictably. We've been demonstrating this on 160m for many years. No respectable professional would dream of putting up a low half-wave for a 2MHz system".

A Practical LF Antenna

The best advice is to put up as much wire as your location, family or the neighbours will allow. In fact you might already have a suitable LF antenna in the form of a G5RV, or 80 or 160m antenna. An 80 or 160m dipole, even if it is fed with coax cable, is a T antenna if the inner and outer conductors of the coax are shorted out.

Any amateur radio LF antenna will nearly always be electrically short; a quarter wave antenna at 137kHz is 550metres (1800ft) long. For this reason you will need a fairly large loading coil to resonate it; in the same way that HF mobile antennas have to use a loading coil.

Loading coils are described later in the chapter. For the time being we will assume that you have a wire antenna and a loading coil, with some means of varying its inductance. In addition, we will assume that you have a multi-tapped transformer and a working transmitter.

Feeding and Tuning

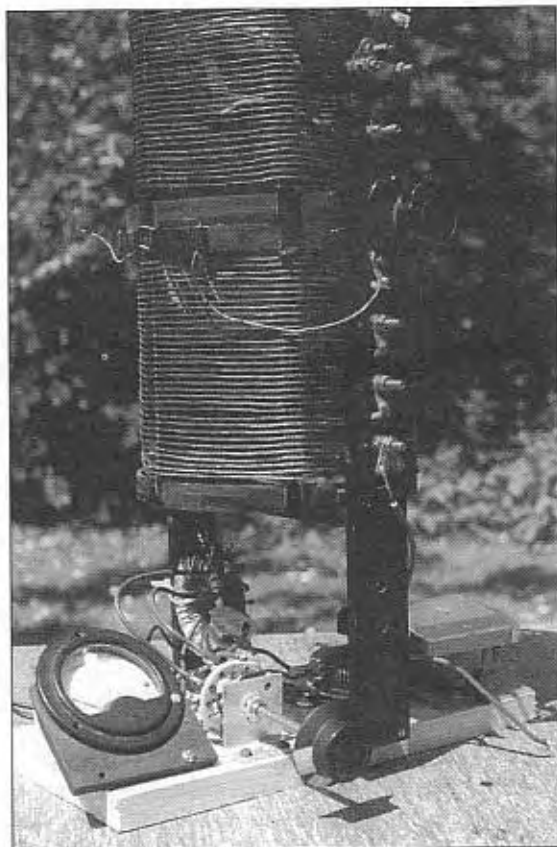
There are two main objectives:

- To resonate the antenna with the loading coil.
- To match the resonated antenna to the transmitter.

An LF antenna setup is shown in Fig 4.6. You can get some idea of antenna resonance using a receiver and altering the multi-tapped coil and variometer for maximum signal or background noise. Most transmitter circuits will not output power unless they 'see' an impedance that matches their output impedance. Some transmitters have been known to be a bit sensitive about their loads and go into oscillation or blow FETs. For this reason low power is advisable at this stage. Transmitter power can be reduced by using lower power-amplifier supply voltages.

The feed systems from the antenna to the transmitter-receiver, used by LF operators, are often 50 ohms impedance. The advantage of this is that you can use slightly modified SWR meters and there is even the ScopeMatch, see Chapter 6.

Fig 4.7: Matching and loading arrangement at G3LDO. The matching transformer is a 3C85 core with windings to match the 40ohm output of the transmitter to the 45ohm impedance of the antenna. The switch is used to alter the secondary winding to suit variations in antenna impedance caused by weather conditions. The meter on the right measures feed current and the one on the left antenna current.



One of the most important instruments for measuring what is going on in your antenna is the RF current meter. You can use a thermo-couple meter but this has the disadvantage of a non-linear scale, with the scale cramped at the bottom end, where you need sensitivity the most when trying to tune up an LF antenna for the first time.

To sum up, you need a current meter to measure the

amount of power going into your antenna and some means of checking that your antenna is matched to the transmitter, as shown in Fig 4.7.

To begin with, check the frequency is clear. Initially, you will be looking for small antenna current readings in the mA range. Use the current meter described in Chapter 6. Alter the antenna loading and the feed matching for maximum current in the antenna.

Once you have established that the loading and matching are nearly correct you can increase the transmitter power and go for maximum antenna current.

Improving Antenna Performance

You have to remember that the LF antenna is very lossy. The reason for this is the very low radiation resistance of a very electrically short antenna, usually in the region of around 0.05 ohms. When you consider this in series with a coil resistance of, say, 5 ohms and a ground resistance of 50 ohms not much of your transmitter power is radiated.

The business of trying to improve antenna performance can be quite disheartening. The obvious area for improvement is the ground connection. However, it is often the case that adding additional copper pipes in the ground seem to make very little difference to the antenna current at all.

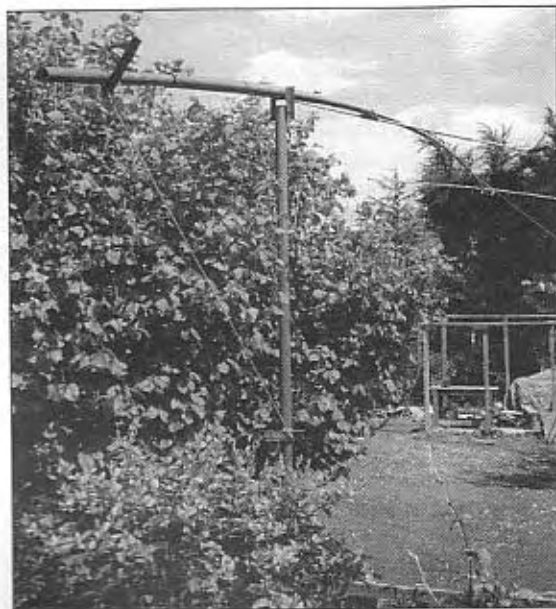
The secret is to work at everything:

- Make sure that the coil is wound so that no wires overlap
 - Add as many ground pipes as far apart as possible
 - Make sure the power amplifier is correctly loaded (this changes with weather conditions)
 - Cut away any unnecessary foliage (I gained an extra 300mA antenna current this way)
- Every little counts!

An Inverted L/V antenna for LF

By Peter Dodd G3LDO

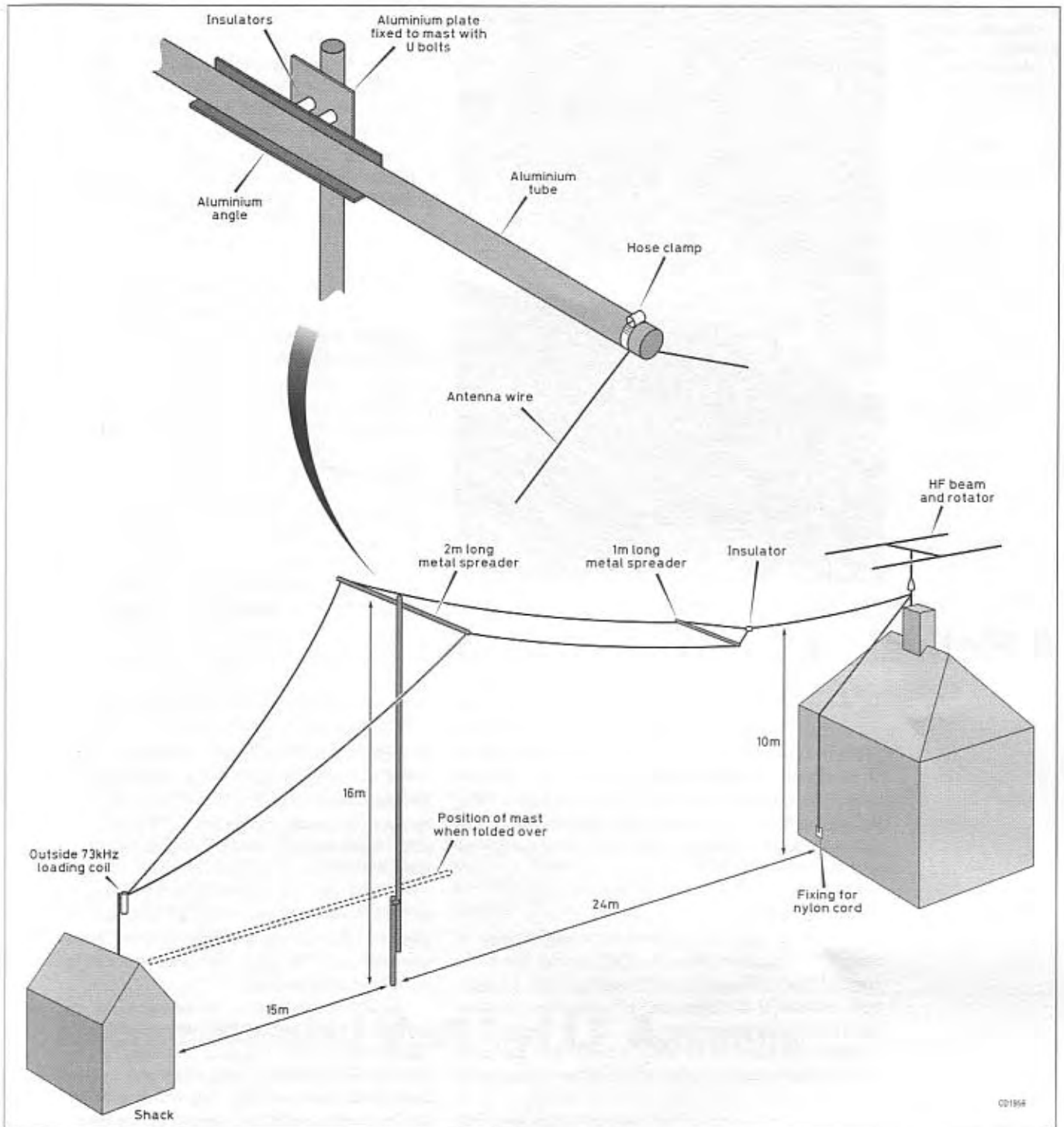
Fig 4.8: Counter-weighted fold-over 16metre mast. The counterweight is provided by using solid steel two-metre long 100mm rod as a section of the lower end of the fold-over section. When the mast was first put up I used a tree trunk as a counterweight. One of the neighbours asked if it was some form of trebuchet for throwing rocks at the houses of antenna objectors!



This antenna just evolved. In the early days of 73kHz I didn't think it would be possible to put up an effective antenna, based on the conventional wisdom of the time. The only mast I have is about two thirds of the way down the garden and it was not possible to erect the traditional capacitive top T or inverted L. Nevertheless I tried a sloping wire from the garden shed, which was used in these early LF experiments. I had tried out some transmission tests with G4JNT on 73kHz using 20W, without success. At that stage I might have given up and then I received a report, via the phone from a local radio ham, saying that I could be heard 4km away. This changed everything.

The position and the design of the mast determined the final antenna configuration so an explanation of the mast is in order. My QTH is located in a private estate and amateur radio antennas are not popular. However, after a battle, I managed to get planning permission for a 16m high mast, provided it is folded over during the hours of daylight!

To manage this restriction conveniently I use a home made fold-over mast, see Fig 4.8, which is counter-



weighted with approximately 15kg (30lb) of top weight so a winch is not required. It takes about 15 seconds to raise the antenna mast into the vertical position. The mast is relatively lightweight; the top third of its length is 5cm (2in) diameter scaffolding pole.

Prior to the advent of LF the mast had been used over a period of time to support a variety of HF beams, and was being used to support a multiband skeleton slot at the time 73kHz was introduced.

Two wires were connected from the shack to the top of the mast. This improved the antenna current and at last I had that elusive 73kHz QSO with G4JNT.

The antenna wires were extended toward the house. Each extra wire added to the antenna system improved the antenna performance in spite of the fixing point at the chimney being six metres lower than the top of the mast.

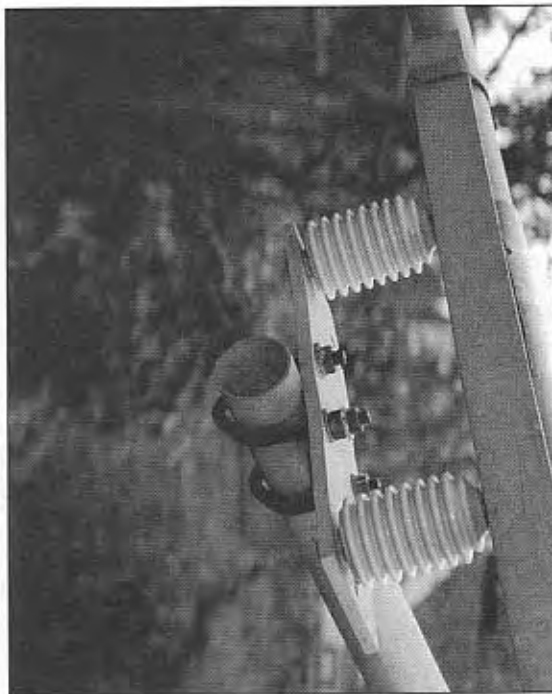
The antenna now comprises 90m of wire configured as shown in Fig 4.9.

To allow the mast to be folded over the fixing at the house end has to be slackened off. The length of the wire from the mast to the house is arranged so that it is well clear of the ground when the antenna is folded over. The wire at the shack side of the mast can be gathered up and draped over a nearby apple tree. It takes approximately two minutes to raise or lower the antenna.

Originally the aluminium spreader at the top of the mast was fixed directly to the mast, with the antenna wires at the end supported via insulators. The antenna wires are now fixed directly to the spreader and insulated from the mast as shown in Fig 4.10. The idea was to increase the amount of metal in the air, although any improvement has not been noticeable.

Fig 4.9: The inverted L/V antenna at G3LDO. The antenna has to be folded over during the day so a pulley is fixed to the chimney to allow the antenna fixing at the house end to be released.

Fig 4.10: Detail, showing the spreader/mast ceramic insulators.



When the hole in the ground for the mast foundations where originally dug out, some copper braid was laid down before the concrete and rocks were put in place. This turned out to be as good a ground system for LF as any. I have since put in copper pipes and a mat of chicken wire on the lawn but none of these moves made much difference to the antenna current.

One of the greatest improvements to the antenna current was achieved by cutting back a lot of foliage growing along the bottom of the garden, although there is still a lot left. See background in Fig 4.8.

A Helical Antenna for 136kHz

By Toni Baertschi HB9ASB

This is a description of a vertical helical antenna, which comprises several hundred turns of wire wound on irrigation tubing and a fishing pole, and uses an existing top band antenna as a capacity top. This antenna is very successful and has been in operation for more than a year now.

Construction

The antenna design was inspired by seeing sections of orange PVC irrigation pipes in a DIY market (the birthplace of many LF antennas and loading coils). I bought three sections of this pipe, each of which were 2m long and 11cm in diameter.

Later a 1m piece of the same material was added, so now the whole irrigation pipe section of the antenna is 7m long.

The upper part of the helical antenna is a 9m long fishing pole from Kevlar. It is fixed one metre inside the top of the irrigation pipe section and the total length of the vertical section of the antenna is 15 metres. The reason for the fishing pole section was to reduce the top weight, which otherwise would present a difficult handling problem: Even so, the irrigation section, with all the wire wound on it, was very heavy and difficult to erect.

The irrigation pipe section of the helix was wound from 5mm diameter Litz wire and the fishing pole section from 2mm diameter Litz. These diameters include the insulation, which is of unknown type.

All together 350m of wire was used in the construction of the helix but the actual number of turns is unknown. Winding the wire on to the PVC tube was difficult, even with the help of my children. The winding was done in packets of about 10 turns close together and with a distance of a few centimetres between the pack-

ets, although the layout of the winding is not critical.

The top of the antenna is connected to my topband antenna. This antenna is 60m long but is constructed in a zigzag to fit into the garden plot of 30 x 40m. The antenna is 18m at one end, fixed to an aluminium tube pole and about 8m high at the shack end.

To construct such an antenna to hit resonance exactly on 136kHz would be impossibly difficult and the self-resonance of this antenna is around 170kHz. The antenna is resonated and matched to 50Ω coax with a separate coil at the base of the antenna.

This coil is wound on a medium sized, plastic wastepaper basket with a 100 or so close wound turns of 5mm diameter Litz wire. The measured inductance is 1mH. This coil is grounded at one end and the coax connected some turns above ground. This coil should have many tapping points near the top to achieve resonance and at the bottom to get the best impedance match.

The advantage of this construction compared with a conventional loading coil is that the voltage gradient is spread over a large distance. The construction has stood the test of time and weather. All winter the coil was completely buried in the snow and certainly soaked with water. No difficulties have been experienced with resonance or loading even with 500W of RF.

The coil and the whole antenna are lightly coated grease, then covered with plastic electrical tape – this used a lot of tape! The lower part of the antenna and the loading coil is shown in Fig 4.11.

Performance

Although the helical works very well, I had always some doubts if this was really worth the effort and if a Marconi antenna at the same place, and the same height, would

work at least as well. In discussions with friends interested in LF we considered the following questions:

- Where does most of the radiation come from; the coil or the top load?
- Does the additional loss of the 350m length of wire, wound on the vertical part, consume the additional dBs obtained from the helical principle?
- Would increasing the helical diameter improve the performance?

The only way to find out how the helical antenna performs was to make a direct comparison with a more conventional Marconi antenna – whose properties are well documented – using the same ground system and of the same height. So last summer I decided to set up the same antenna configuration just a few metres away and with a straight wire instead of the helical part. To make it fair comparison a huge loading coil was built on the largest plastic barrel available from a garden centre. In addition the coil was wound using 5mm Litz wire, the same as used for the lower part of the helical antenna.

The same top load (the zigzag 160m antenna) and the same ground network were also used.

The second antenna worked very well and the match was perfect. However, the helical was at least 3dB better than the Marconi. Not only did it produced a stronger transmitted signal but the received signals were also 3dB greater when using the helical antenna. In addition, the bandwidth of the helical antenna is wider so there is no need to retune the antenna during a QSY to a different frequency.

Extensive tests on this antenna included ground wave tests with HB9DCE, 140km away, and many longer distance contacts during day and night times. Field strength measurements, performed using a Wandel & Goltermann EMR-20 FSM, confirmed that the helical antenna was least 3dB up on the Marconi.

The answer to the first question above is that a large proportion of the signal must be radiated from the helix. As regards the second question, if there are resistive losses associated with the 360m of wire in the helix they are



Fig 4.11: General view of the lower section of the helical antenna. The winding packets can be clearly seen. The plastic waste-paper basket loading coil can be seen in the foreground.

more than compensated for by the design of the antenna.

With regard to the third question – that has yet to be resolved. Lately I've discovered much larger PVC tubes in another DIY market; these comprise long 6 metre sections, with a diameter of up to 50cm. Maybe I will try another larger helical antenna.

I think it is worth the effort – not for the 3dB of course, as this is easy to compensate for with Tx power, but it shows us that the common Marconi T-antenna is not the final answer to our antenna problem on longwave, and it could be a good solution for a small garden and other antenna restrictions. If you can handle larger tube diameters the helical is good for some of those extra elusive decibels!

Balloon-extended Vertical LF Antenna

By Steve Rawlings GW4ALG

I have a small suburban plot so a large LF antenna structure is out of the question. To overcome this restriction I have designed a lightweight structure whose height can exceed 20 metres in the right weather conditions. The antenna can be quickly erected or dismantled.

The main part of the vertical uses a 10 metre long fibre-glass pole with 5 metres of a 22 metre length of insulated wire taped to it as shown in Fig 4.12.

The base of the glass fibre pole is mounted on a wooden base assembly, which slides on to a wooden upright, fixed to the frame of a children's garden swing. (As our children became older, the homemade swing frame had fallen into disuse – until now!)

The top wire section of the antenna can be supported using balloons. I had considered just using the balloons and dispensing with the pole. However, although a balloon can be used to support any wire antenna on its own

there is one big disadvantage. If there is any wind blowing at all the balloon tends to be blown sideways so that the angle of the antenna relative to the ground is considerably less than 45 degrees. Not only does this severely reduce the height but the antenna is in grave danger of becoming entangled with nearby buildings and obstacles.

One option is to use a balloon with an aerodynamic shape so it doesn't get blown sideways. This method is used with balloons for advertising, and of course barrage balloons during the war. However, advertising balloons are very expensive so a different solution was sought.

It occurred to me that it would be best to use balloons just to extend the existing vertical. This would mean that the balloon would have to carry less wire. Also, with the tethering height of the balloon 12 metres high the chance of it becoming entangled with nearby obstacles would be reduced considerably.

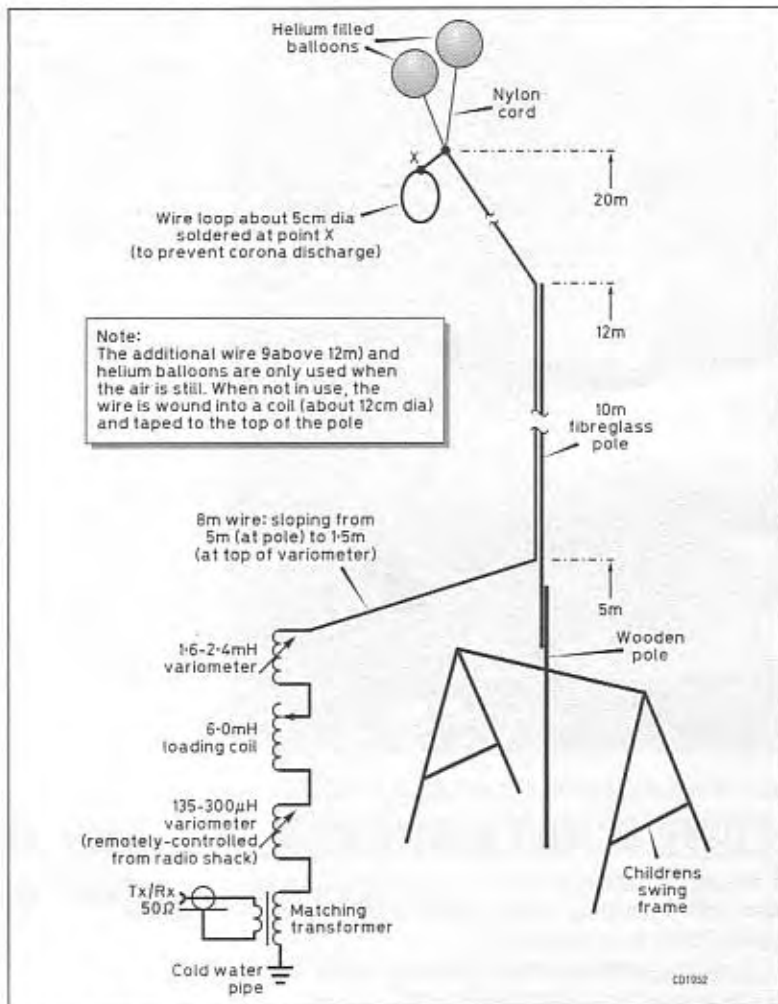


Fig 4.12: The GW4ALG balloon extended vertical. The extension wire and balloons can only be used when the wind is very light. When the extension is not in use the 8 metre length of wire is wound into a 12cm coil and taped to the top of the pole.

Additionally, if the wind is too strong to use the balloons the vertical can be used on its own.

This free-standing vertical section is not strong enough to withstand high winds. So this is definitely a 'fair weather' antenna, for the time being at least! The anten-

na is taken down at the end of each operating period.

At the top of the wire, a loop of about 6cm in diameter has been made by folding the wire back on itself and soldering it in position. The purpose of the small loop of wire is to help prevent corona discharge from the top of the vertical.

Balloon information

I use 'Punchball' type balloons, as used at children's parties and each balloon costs about 40p. The balloon can be inflated to around 400 mm diameter and can be re-used about 3 times, if not over-inflated.

The neck of the balloon may be folded over and held with a small cable tie to seal it.

Two 400mm balloons are used to lift 8m of insulated, multi-strand wire. An insulating section comprising at least 300mm of thin nylon cord (or thick kite string) is required between the top of the wire and the balloon. This is to prevent arcing - a 400mm balloon makes a very big bang when it bursts and it took three 'bangs' before I figured out what was going on.

Each 400mm balloon will cost about 90p to inflate with gas; much of the helium gas leaks out of the 'Punchball'-type balloon within 8 hours because helium is lost through the membrane of a balloon at a faster rate than air molecules.

Consult 'Yellow Pages' for balloon suppliers (who will also be very willing to supply the balloon gas).

Operation

With the basic 12m antenna I was able to work OH1TN (1916km), which at the time set the world DX record on 136kHz, and remained the UK distance record for over 12 months.

At the time of writing, I have worked 13 countries using the basic 12m antenna, these being: DL, EI, F, G, GD, GI, GM, GW, HB9, OH, ON, PA and SM. I went on to make several more QSOs with Reino using both the 12m and 20m (balloon-supported) set-ups and have now completed 23 two-way CW QSOs with various stations at distances exceeding 1000km.

I really hope that these results will encourage others with small gardens to give 136kHz a go.

Limited Space 136kHz Antenna

By Rik Strobbe ON7YD

This antenna is an inverted L with a four wire capacitive top, 11m high at the ends with around 1m sag in the middle. The length of the antenna is 23m and the width is 2.4m (four parallel wires spaced 0.8m apart).

The antenna is supported by four 15m high birch trees. This presented the usual challenge of trying to maximise the efficiency of the small antenna. Additionally there was the mechanical challenge of designing the antenna to survive strong winds. Originally the rope kept breaking because of friction caused by movement of the trees in strong winds. The problem was solved by inserting bungee cord in the system as shown in Fig 4.13. This material is only about 0.8m long in the normal state is but it can stretch to over 2m without breaking. Additionally

some short pieces (about 3m) of PVC coated steel rope were placed where friction with the branches occurred. The antenna has now been up now for over a year and has withstood many strong winds without damage.

Regarding the antenna efficiency, the main problem is high loss resistance. When I started on LF in early 1998, I measured a loss of 220 ohms! Gradually the ground network was improved. Initially this consisted of six short radials (10 to 20m each) in the ground and was extended to a complex system of six ground rods (galvanised steel, 2.5m deep in the earth) and about 150m of galvanised iron wire interconnecting all the rods and buried about 20cm deep in the ground. But even with this ground system the loss resistance is still 120 ohms. Further exten-

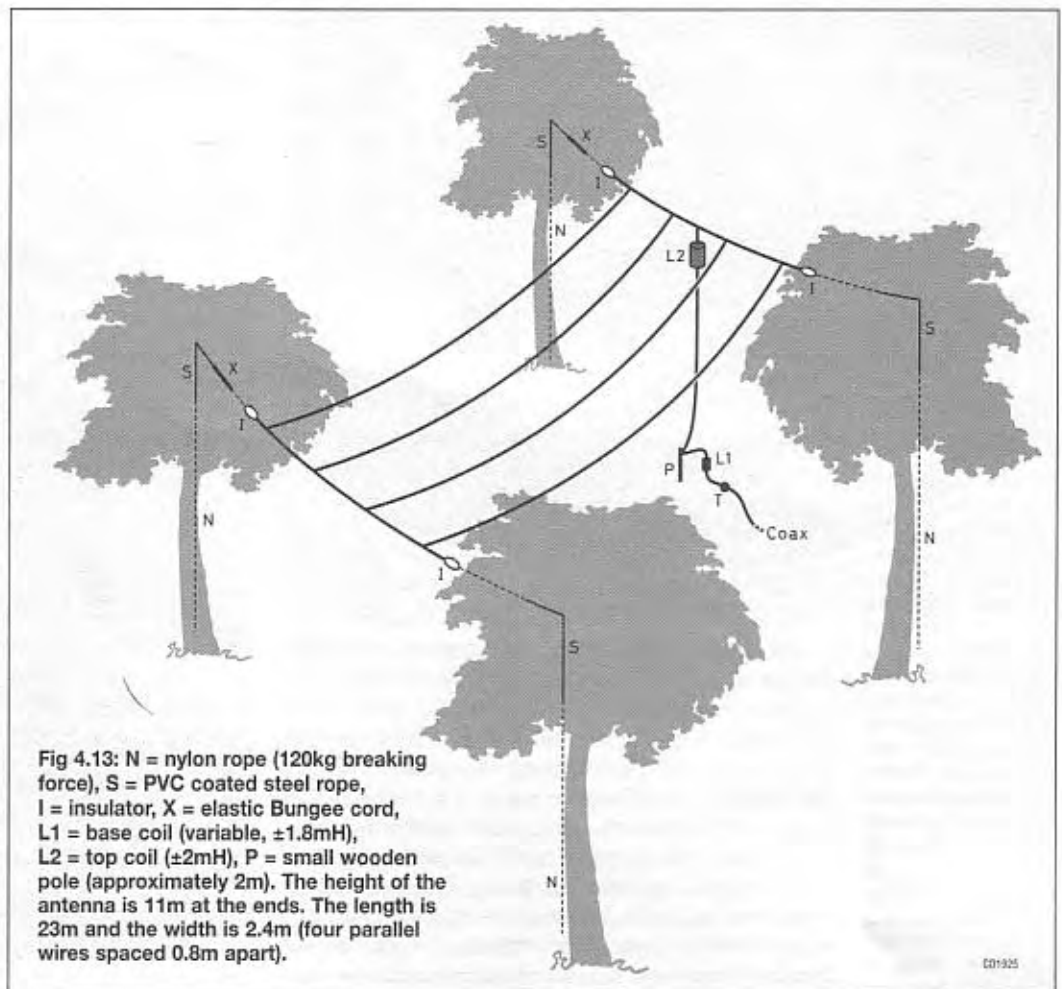
sions of the ground system gave little or no reduction of the loss, which leads me to believe the loss is due mainly to the several hundred trees surrounding the antenna, rather than ground loss.

An unexpected improvement was achieved by placing about 50% of the loading inductance at about 10m from the ground. The expected gain (calculated and simulated) was only about 0.5dB but several tests and measurements showed an improvement of 4 to 5dB. A possible explanation for this unexpected gain is the fact that the voltage on the vertical part is reduced by about 50% while the top load voltage remained almost unchanged. This would mean that the unwanted 'coupling' of the antenna to the many surrounding (vertical) trees comes mainly from the vertical part of the antenna and by reducing the voltage on the vertical part this loss is also reduced. This increase in gain was achieved without any increase in antenna current.

In fact the current has dropped slightly (from 1.9 to 1.85A) probably because of the extra loss in the elevated coil (because this coil had to be very light weight I used thin wire).

With about 450WF power I get an antenna current of

about 1.85A. Estimated ERP is about 100mW. But despite this very modest ERP I have managed to work over 70 stations in 17 DXCC countries during the past 2 years. This might encourage other amateurs to become active on LF, even if they have limited space.



The Loading Coil

By Peter Dodd G3LDO

Because most LF antennas are electrically short you will probably need a loading coil of around 3 to 4mH. Finding coil formers of the right size for such a large inductor can be a problem, so various solutions are described here. It is often difficult finding a plastic tube with a large enough diameter, so a bucket or a bin is often pressed into service for this purpose. Examples of LF inductors, wound on different types of former, can be seen in Fig 4.14. Most experimenters have found that 1 to 1.5mm plastic insulated wire is suitable for constructing LF loading coils.

The voltage developed across an LF loading coil is very large. Because of this the coil must be wound as a single layer solenoid so that the RF potential gradient is distributed over the full length of the coil. Any attempt to increase the inductance by using multi-layer construction will probably result in flashover.

Other methods of making coil formers include using



Fig 4.14: Three examples of loading coils by G3KAU. On the left the coil former is a large diameter plastic pipe. The centre coil is wound on a plastic rice container while the one on the right is wound on a plastic bucket. (Photo 14)



Fig 4.15: A coil wound on plastic lattice fencing. There is very little plastic material used in the former so the coils wound on them are almost air-spaced. The coil is wound in small multi-layer bunches distributed over the full length of the former and held in place with hot melt plastic adhesive.

plastic sheet or plastic lattice fencing rolled into a cylinder. An example of a coil wound on plastic lattice fencing is shown in Fig 4.15.

A further requirement is for the inductance of the coil to be variable, in order to resonate the antenna at the correct frequency. This is often achieved by using two or more coils in series. One of the coils can be tapped for coarse tuning and the other can be a variometer for the fine-tuning. An example of an arrangement using a fixed inductance in series with commercial variometer is described in 'A guide to LF Antennas', earlier in this chapter. There are a number of these commercial variometers around and although most of these are designed for around 500kHz they are useful when used in conjunction with a larger fixed inductance.

Making a variometer is not as difficult as it might first appear. An example of a home-made variometer is shown in Fig 4.16. By making the fixed coil with a large diameter former and the moving coil with a small diameter former the mechanical problems of having one coil rotate inside the other are overcome.

Details on how to design a variometer are described later in the chapter.

The coil former designed by PA0SE uses a different approach to both overcoming the large diameter coil former and the tuning. The coil former consists of four 1m

Fig 4.16: A variometer made by GW4ALG. The fixed coil is wound on a plastic bucket and the inner moving coil is wound on a section of plastic pipe. The spindle is turned from a section of wooden dowel.

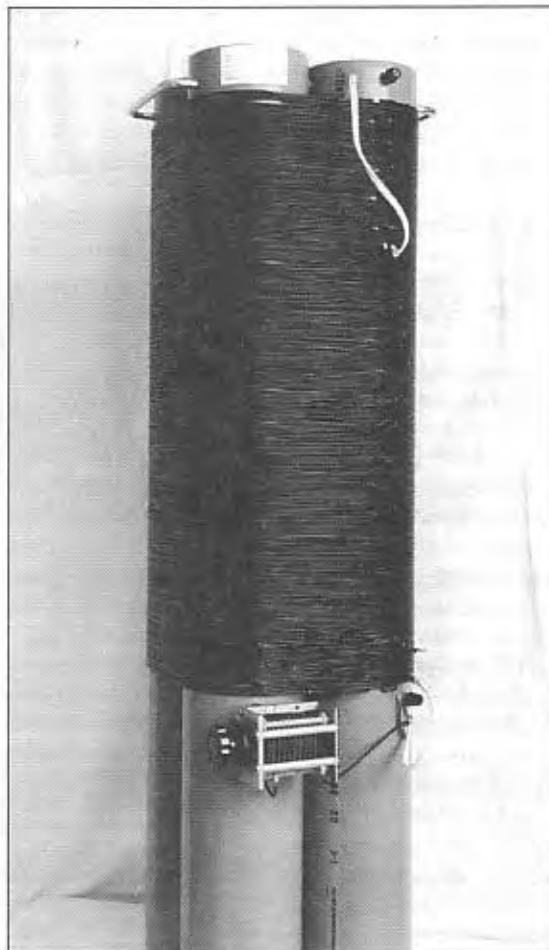


Fig 4.17: PA0SE's antenna loading coil for the 136kHz band. Just visible is the capacitor used for fine-tuning. A thermocouple RF ammeter is used to measure antenna current.

long PVC tubes of 12.5cm diameter (5-inch) bolted together to form a square former with rounded corners, as shown in Fig 4.17.

The coil has 200 turns of PVC insulated house-wiring wire of 1.4mm diameter (17 SWG). It is close wound with taps at every 10 turns at the top end and every single turn for the first 12 turns from the bottom. Total inductance is 4.4mH. By short-circuiting turns at the top and bottom end, inductance can be decreased in steps of one turn. The measured Q is about 350.

Fine-tuning is achieved using a vacuum capacitor in parallel with the coil. With proper selection of the coil inductance the value of the vacuum capacitor can be kept near its minimum. (The current in the capacitor causes extra current in the coil, increasing its loss.) The transmitter is connected to one of the taps near the bottom end of the coils, selected for the best combination of antenna current and transmitter loading.

A Lightweight Loading Coil

Some experimenters (G3XDV and ON7YD) have found that antenna performance can be achieved by using an elevated coil as part of their antenna system.

Most of the radiation takes place from the element below the loading inductance, so the coil should be in some elevated position. The problem with a wire antenna is how to elevate coils. They can be supported by the wire element itself but a good high-Q coil can be fairly heavy.

The following description of the coil is the result of

experimental work by Mike Dennison, G3XDV.

The coil former is made from a large two-litre size plastic soft drinks bottle. The plastic used with these bottles is rather thin and is normally held in shape by the pressure from the gas in the soft drink. When the bottle is empty it is rather floppy and no good as a coil former. The trick is to pressurise it. This is done by simply placing the bottle in the freezer compartment of a refrigerator for a couple of hours without its top. When the bottle is taken out of the freezer the top is screwed on firmly. As the air in the bottle warms up it expands and pressurises the bottle, which then makes a good lightweight coil former. The wire can now be wound around the bottle. The wire is held in place using hot-melt glue and finally fixed using plastic cement. A complete loading coil is shown in Fig 4.18.

The loading coil can then be weatherproofed by coating it in clear silicone rubber.

How Many Turns

The big question, particularly for the newcomer to LF, is how big should the coil be? For most practical LF antennas that can be accommodated in the average back garden, an inductor of between 3.5 and 4.5mH is required. If we use a plastic bucket, for example (use white plastic – the chances are the losses will be less), of around 250mm (10in) diameter then for an inductance of 4.4mH you will need 160 turns of plastic covered 1 to 1.5mm diameter wire. Plastic covered wire has the benefit of being cheap and easy to obtain and it imposes a wire spacing that gives a reasonable Q. There is little evidence of increased losses at these frequencies using this type of wire.

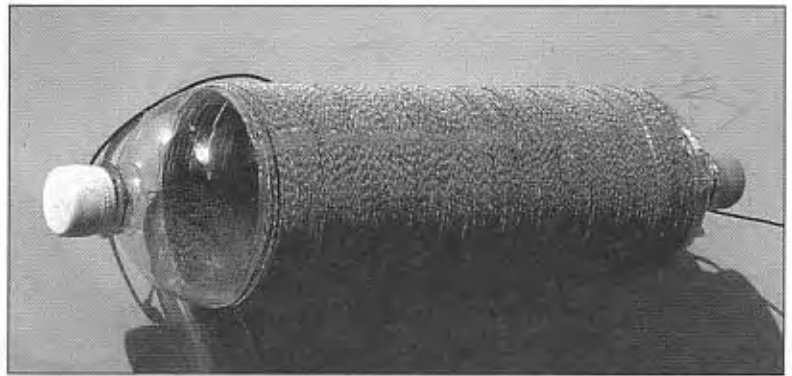
This will require 126 metres of wire. The tapping arrangement described by PA0SE is recommended.

A program for calculating the inductance has been written by Reg Edwards G4FGQ. A screenshot from this program is shown in Fig 4.19. Once the program has been run using an initial set of parameters, the inductance and effective series resistance is noted. The coil length and number of turns can be altered for the lowest loss for the coil former that you are using. See the Appendix for the availability of this software.

Basket Weave Coils

The time-honoured method of reducing the size of the coil and the coil self-capacitance is to use basket weave winding. This skill is all but forgotten but can still be found. *The Wireless World and Radio Review* has a very extensive article on the subject. Some of the designs are very convoluted.

Lyle, K0LR gives some brief notes on building these coils for those who wish to experiment. A basket-wound coil of any size can be constructed by putting an odd number of dowels into holes in a plywood board in a circular configuration. The PVC insulation provides the right wire spacing so it isn't too difficult to wind. After 'weaving' the coil, lace the crossovers with something like Dacron kite string, and dope the lacing with silicone bathtub sealer. You can then pull the whole thing off the dowels and have a reasonably efficient loading coil at low cost. If your coil is heavy it could probably be left in place to help keep the coil rigid without too much degradation of Q. Johan Bodin, SM6KLM says that he measured a Q of over 400 on his basket-wound coil. The basket-wound configuration is probably worth the extra effort when using wire with perceived lossy insulation.



Some of the best examples of highly efficient basket wound coils are those from the decommissioned Decca transmitter stations as shown in Fig 4.20. These are wound using spacers to give the smallest possible coil capacitance and reduce proximity effect to a minimum.

G3XDV, connected up one winding of one of these Decca coils, and the antenna (plus the coil at the top of the mast) resonated at 150kHz. Connecting another in series reduced the frequency to 88kHz. This is not the sort of coil you can tap so it was back to the old dustbin supplement, a single Litz coil. At the time of writing, using three windings to get to 73kHz had not been tried. The built-in variometer changes the frequency by only a few kHz.

The antenna current showed a 20 per cent increase compared with the solenoid coil wound with PVC insulated wire so the coil loss is obviously much reduced. In the early experiments with LF there was a lot of discussion about Litz wound inductors. The conclusion then was that any advantage would be small compared with earth and other losses. However, it does show that once you have reduced the other losses as far as possible there is some advantage in going for the lowest loss coil available.

If you have a very small antenna you will need a larger inductance coil. To get more wire on to a given size former you can use enamelled copper wire rather than PVC insulated wire. An example of such a coil can be seen in the description of 15TGC's station in Chapter 1. An example of an even larger coil is shown in the Tesla coil, also in Chapter 1.

Litz (endraht) Wire

Most of the examples of coil building described use PVC covered electrical wire. Commercial stations use large diameter coils wound with Litz wire cables, comprising of hundreds, or even thousands of insulated strands, each of 0.1 - 0.25mm diameter. By using separate insulated

Fig 4.18: A pressurised plastic bottle used as a coil former. The coil is fixed to the top element of the antenna using plastic tape around the neck of the bottle. An additional 'neck' is cut from another bottle of the same size and glued to the bottom of the coil former. This forms a fixing point for the bottom wire element of the antenna. This coil is wound with thin Litz wire by G3LDO and has an inductance of around 2mH.

Fig 4.19: Results of a coil calculation using the G4FGQ Solenoid program. Any parameter may be changed, such as the number of turns or the length of the coil, and the effect on the inductance and the effective series resistance noted.

SOLNOID2			
Add			
L. Length of coil, mm	415.0	R. Ratio wire Dia/Pitch	0.950
D. Diameter of coil, mm	240.0	F. Test Frequency, MHz	0.137
N. Number of turns	200	V. Applied RMS volts ..	1
Coil VLF Inductance 4419.502 microhenrys			
Coil impedance magnitude 3891.8 ohms, includes self-capacitance			
Impedance phase angle 89.84 degrees			
Effective series resistance 10.362 ohms, all losses combined			
Q in a parallel tuned circuit 367 approx, at test frequency			
Q in a series tuned circuit 359			
Self-resonant frequency 0.914 megahertz when coil is isolated			
Self-capacitance 6.87 pF when coil is isolated			
Additional tuning capacitance 298.50 pF resonates coil to test freq.			
Resonant impedance of tuned ckt 1398.72 k-ohms, at test frequency			
Wire diameter 1.98 mm = 0.078 inches			
Winding pitch 2.09 mm = 0.082 inches			
Length of wire 152.04 metres = 498.8 feet			
Power dissipated in coil 0.00 watts, at test freq and volts			
Vary Ratio with keys 1,2 Length: 3,4 Diameter: 5,6 Turns: 7,8 Freq: 9,0			
Hit keys L,D,N,R,F,V to change data. B(egin again) or Q(uit) ...			

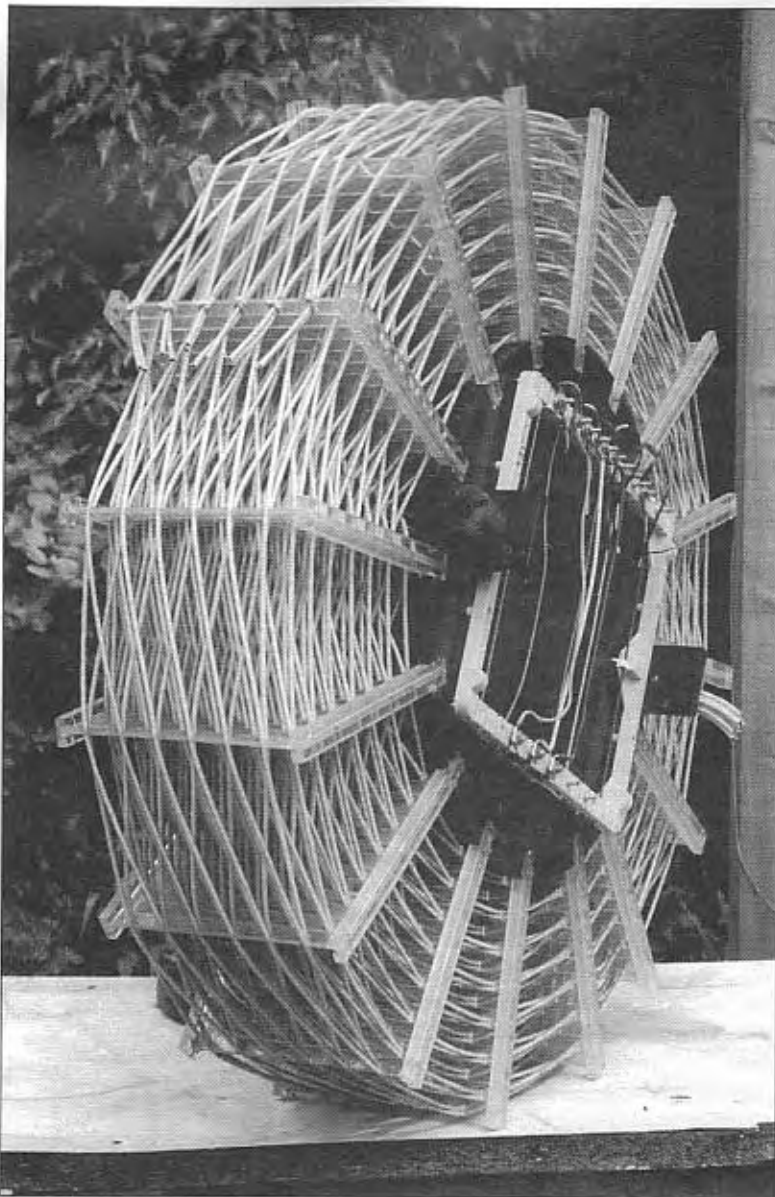


Fig 4.20: An example of highly efficient basket weave coils. These ex-Decca loading coils have four parallel-wound windings - each wound with 4mm, 729-strand Litz wire. The coil has a built-in variometer for fine tuning.



Fig 4.21: 'Cooking' Litz wire in an electrically heated solder pot. When the burned off insulation floats to the surface of the solder all the conductors are tinned.

conductors in parallel the relative weighting of skin effect (concentration of high frequency currents in the surface of the conductor) are considerably reduced. Coil Q-factors of 1000 to 3000 can be achieved for inductances up to 10mH using Litz wire and coil construction methods shown in Fig 4.7. Litz wire is not useful for frequencies greater than 500kHz.

The most important consideration when using Litz wire is that *all* of the individual strands are connected, otherwise the benefits of using Litz wire are lost. Modern Litz wire appears to be constructed from individual strands whose insulation can be removed by just heating it up. In practice this is done by using a solder pot. The Litz is covered with a plastic sheath, which can be softened by dipping in the solder pot for a few seconds then stripped off. The conductors are then tinned shown on Fig 4.21:

Not all Litz wire has insulation that can be removed from the conductors as described.

Jim Moritz M0BMU recently salvaged some thick Litz wire, and looked for a workable method of stripping the insulation. The enamel is of the older, dark coloured type, which cannot be removed by dipping in molten solder. The overall conductor thickness is about 4mm with 729 strands, and a spirit burner does not make any impression. A blowtorch does, but also this also burns away the outer strands and the plastic sheathing.

However, the following seems to work quite well - strip about 50mm of the plastic sheathing, and wrap the end 25mm of wire strands tightly in a piece of copper foil (such as can be peeled off some PCB laminate, or the type used for EMC shielding). Wrap a smaller piece of foil around the 12mm of wire strands nearest the stripped back sheathing, and grip with a pair of pliers - this acts as a heatsink to protect the sheathing. Heat the foil-covered end red hot with a blowtorch. The enamel burns off with a small flame at the end of the foil, while the foil protects the copper strands from the blowtorch flame. When the flame goes out, leave to cool and unwrap the foil. The wire strands are coated in black, shiny residue, which can be cleaned off by rubbing the end of the wire against the bottom of a shallow dish full of water, as if cleaning a paintbrush. Somewhat surprisingly, this leaves clean, shiny strands of copper that can be soldered in the usual way.

Low Frequency Coil 'Q'

By Bill Bowers

Over the years I have come across a number of recommendations for coil diameter to length ratio for helical or basket weave type coils that will produce the maximum Q. The problem has been that none of these recommendations was backed up with theoretical or experimental data, nor was there any indication as to the frequency range for which it applied.

In all the amateur literature, including a computer program, the recommendation is to make the coil 'square', or in other words make the coil length (L) equal to the coil diameter (D). Curty Communications also states in their instructions that come with their Lowfer transmitter kit that for maximum Q the antenna coil should have a length equal to the diameter; $D/L = 1.0$.

Dave Raney sent me a copy of several pages from *Antenna Engineering Handbook* which states that "...optimum Q is obtained when the antenna loading coil diameter is twice the loading coil length: $D/L = 2.0$."

Mitchell Lee in his article 'Basic 1750 Transmitting Antenna' states that: "for optimum Q the diameter to length ratio should be $D/L = 2.5$."

With this wide range of recommendations I was in a real quandary as to what I should do to get the best coil, so I decided to run a few tests on different configurations. The tests in this report were all run at 200kHz, which is low enough to apply to the 136kHz band. All measurements were made with a Boonton Model 260-A Q meter. The Boonton Model 513-A Q Standard was used to verify the meter accuracy.

The first series of tests were run using a white 5in PVC

pipe as a coil former. Since the longest coil length suggested was equal to the diameter, the test started with a coil length of 5in and with enough turns to give 0.5mH inductance. The coil turns were then pushed together and the turns reduced to keep the inductance constant. The test results are shown in **Table 1**.

From this test series it appears that Mitchell Lee had recommended the best setup for low frequencies. It also looks like going shorter is much better than going longer.

Using the optimum D/L ratio the next series of tests consisted of changing the wire types. In addition the coil former was cut to make 9 fingers about 1/4in square and 4in long so that a basket weave coil could be tested. The results are shown in **Table 2**.

This test clearly shows the advantage of using as large as possible Litz wire. The surprise was that there was very little gain in going for basket weave construction. This would indicate that the losses caused by distributed capacitance are far less at low frequencies than at higher frequencies where basket weave can make a significant improvement.

One further arrangement was tried which consisted of winding the coil over the f-I cut in the coil form in a standard helical manner. In this setup Q was almost equal to the basket weave coil. This result would suggest that if you do not want to go to the trouble of winding a basket weave coil then cutting slots or even drilling holes in the coil form could improve your Q.

To complete the testing it was decided to test bigger coils of the size that could be used for Lowfer antenna

D	L	D/L	N	Q	WIRE	COIL
5	5	1.0	79	235	100x 44 Litz	Helical
5	2.5	2.0	64	280	100x 44 Litz	Helical
5	2	2.5	60	330	100x 44 Litz	Helical
5	1.63	3.0	57	325	100x 44 Litz	Helical

Table 1: Effect of altering D/L ratio

D	L	D/L	N	Q	WIRE	COIL
5	2	2.5	60	308	60 x 42 Litz	Helical
5	2	2.5	60	330	1 00 x 44 Litz	Helical
5	2	2.5	60	505	200 x 44 Litz	Helical
5	2	2.5	60	525	200 x 44 Litz	Basket weave
5	2	2.5	60	520	200 x 44 Litz	Helical over fingers
5	2	2.5	60	195	22AWG	Solid helical

Table 2: Effect of altering wire type and winding method

D	L	D/L	N	Q	WIRE	COIL
19	9.5	2.0	44	300	50 x 36 Litz	Basket weave
19	7.6	2.5	42	410	50 x 36 Litz	Basket weave
19	6.3	3.0	40	400	50 x 36 Litz	Basket weave
19	4.0	4.7	37	345	50 x 36 Litz	Basket weave

Table 3: Effect of D/L ratio with a large former

D	L	D/L	N	Q	WIRE	COIL
19	7.6	2.5	42	410	50 x 36 Litz	Basket weave
19	7.6	2.5	42	430	200 x 44 Litz	Basket weave
19	7.6	2.5	42	625	600 x 44 Litz	Basket weave
19	7.6	2.5	42	325	14AWG solid	Basket weave

Table 4: Effect of altering wire type whilst keeping D/L ratio constant

loading coils. A 19in PVC sewer pipe was located and cut to give 15 half-inch square fingers that were about 11in long. The same series of tests were run except that the inductance was increased to 1.0mH (Table 3). Since the object of this test was to determine the best D/L ratio for large coil diameters I just used some 50 x 36 Litz wire which I had on hand.

It looks that even for larger diameter coils a ratio of 2.5 is again the best diameter-to-length ratio, and again shorter is better than longer.

The final test was to compare different types of wire on this larger coil using the 2.5 D/L setup which looked best in both tests.

This last test (Table 4) is quite interesting in that it shows that readily available and cheap solid #14 wire produces a fairly acceptable Q when used in a coil with a 2.5D/L ratio. If the ultimate Q is what you are after, then get Litz wire with as many strands of 42 or smaller wire as possible. In the test above the coil with 600 x 44 Litz gave a Q value much greater than 625, but this was as high as the meter could read.

One other interesting test was run which consisted of again winding a 1.0 mH coil but on a 7in diameter basket weave coil form to compare it with 1.0mH coils wound on 19in coil forms using the same 200 x 44 Litz wire. Using 72 turns and a length of 2.8in (D/L = 2.5) the inductance was 1.0mH and the Q was 500. This compares to a Q of only 430 for the same inductance and type of Litz wire when wound on the 19in former. This confirms what I had experienced in previous tests. The rule, therefore, is to use as small a diameter coil former that will accommodate the number of turns required and still maintain the D/L ratio of 2.5. This recommendation would only apply at low frequencies where the effects of distributed capacity are at a minimum.

Further proof of this conclusion can be shown by computing the length of wire (W) required for the 19in and 7in 1.0mH coils:

$$W(19) = 11 \times 19 \times 42 = 2507 \text{ inches}$$

$$W(7) = 11 \times 7 \times 72 = 1583 \text{ inches}$$

This comparison shows that it takes less wire to produce a 1.0mH coil on a 7in diameter than on a 19in diameter with the same optimum D/L ratio. Less wire means less coil resistance and hence the 7in coil should have higher Q, as experiments have shown. Mitchell Lee states in his article on Litz wire back in the April, 1987 *LOW-DOWN* that the maximum inductance for a given length

of wire is obtained when the coil D/L ratio is 2.46. This would further indicate that the effects of distributed capacity are small and the Q of coils at LOWFER frequencies are primarily determined by AC conductor resistance.

So far all of this testing has shown that diameter should be 2.5 times the coil length, and the coil diameter should be as small as possible, but the optimum coil dimensions can be worked out.

l = the inductance required in micro-henries.

d = spacing of turns. If you are using plastic insulated wire then this should be the diameter over the insulation. For bare or enamelled wire then d should be 1.25 to 1.5 times the wire diameter to allow for a space between turns – inches.

D = coil diameter – inches This is the outside diameter of the coil former. For basket weave windings D is the outside diameter less the thickness of the fingers or rods.

L = the resulting length of the coil.

N = the required number of turns.

Once the required inductance has been determined and the best wire located then the other parameters can be calculated as follows:

$$D = 6 \times [l \times d^2]^{1/3}$$

$$L = D / 2.5$$

$$N = L / d$$

EXAMPLE: If you need 2mH resonate your antenna and the wire available is enamelled solid 14AWG, for maximum Q build a coil like this:

$$D = 6 \times [2000 \times 0.080^2]^{1/3} = 14\text{in}$$

$$L = 14 / 2.5 = 5.6\text{in}$$

$$N = 5.6 / 0.080 = 70\text{turns}$$

$$D / L = 2.5$$

$$l = 2000 \text{ microhenry}$$

$$d = 1.25 \times 0.064 = 0.080\text{in}$$

A final cross check is to calculate this coil's inductance using the classic formula for single layer helical coils as given in the *ARRL Handbook*

$$l = [D \times N]^2 / [18D + 40L] D = 14\text{in}$$

$$l = [14 \times 70]^2 / [18 \times 14 + 40 \times 5.6] \quad N = 70$$

$$l = 2017 \text{ microhenry}$$

$$L = 5.6\text{in}$$

There is no simple way of calculating the actual Q for this coil, but based on the above experiments this configuration will give the maximum Q possible for this inductance and wire size in the 1750 metre band.

LF Variometer Design

By Tony Preedy G3LNP

The traditional variometer of the 1920s crystal receiver when appropriately rated is a convenient means of tuning an LF transmitting antenna.

Construction is straightforward using 6mm shafts running in bushes salvaged from old volume controls and doubling as terminations for the flexible "pigtailed" forming inner to outer coil connections as shown in Fig 4.22.

These mechanics are compatible with standard insulated shaft couplers and control knobs. The important considerations are to make sure all the windings have the

same direction and that rotation is restricted to 180 degrees. It is advisable to use a working tap that puts the inner coil in the additive inductance position to minimise losses.

The following data is based on practical construction experience.

Dimensions

The normalised dimensions shown for a variometer type antenna tuning inductor in Fig.23(a) will allow an induc-

tance (L) range of 6:1 in three stages without gaps. Taps 1, 2, 3 (Fig 4.23b) provide respectively $2L/3$ to L , $L/3$ to $2L/3$ and $L/5$ to $L/3$.

Unused turns below tap 2 can be left open whilst unused turns below tap 3 should be short-circuited to minimise both voltage stress and losses.

Design Method:

- Decide on the antenna current, maximum inductance and acceptable Q. The formula below gives the maximum wire size for close winding. Choose the minimum solid wire size on the basis of 5 Amperes RF per mm diameter for typical amateur duty cycles. In either case wind the inner coil with turns touching to maximise the incremental inductance. The range of inductance variation can be increased by inserting iron dust or ferrite material inside the inner coil although it is not necessary with this design [Beware – ferrite rods from broadcast radios can get very hot, better to use ferrite cores from SMPSUs or line output transformers]. Litz wire size is calculated by dividing 5 Amps between the strands (Litz wire will carry more RF current for a given diameter and results in a smaller assembly for a given inductance). Choose a convenient diameter D to start your calculations and then adjust this until you have an acceptable Q as follows:

- Use the formula below to calculate the turns on the larger coil for maximum inductance L (ignore the small coil).

$$N = (50L / D)^{1/2}$$

where D is diameter of outer coil in inches and L is in μH .

- Use this formula for maximum wire diameter in mm:

$$d = 25D / n$$

- Q can be roughly estimated by assuming that RF resistance for solid wire will be 10 times DC resistance. Spacing the turns will reduce this factor. For close wound Litz wire I have found the RF resistance to be typically 3 times DC resistance at 136kHz.

Total wire length (both coils) in metres is:

$$l = nD / 10$$

DC resistance $R = l / m$ per ohm where m per ohm is taken from copper wire tables.

Examples.

A 3mH inductor with a Q of 400 is required to tune a T connected G5RV doublet to 136kHz:

(I know from experience that this requires a diameter of 8 inches so I will start with $D = 8$)

Length of former = $1.25 \times 8 = 10$. Allow an extra 2in for terminals and fixing.

$$\text{Turns } n = (50 \times 3000 / 8)^{1/2} = 137.$$

Wire size $d = 25 \times 8 / 137 = 1.45\text{mm}$ (nearest SWG is 17 with a length per ohm of 170m).

$$\text{Wire length } l = 137 \times 8 / 10 = 110\text{m}.$$

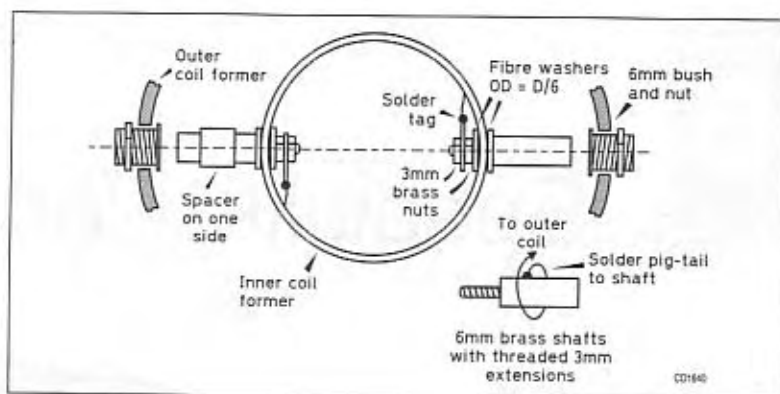
$$\text{DC resistance } R = 110 / 170 = 0.64\Omega.$$

$$\text{RF resistance } R_{rf} = .64 \times 10 = 6.4\Omega.$$

$$Q = XL / R_{rf} = (6.28 \times 136 \times 3) / 6.4 = 400.$$

Examples of Q measured at 136kHz on variometers built to this design:

- 15 Amps rating using 2mm diameter Litz for 500 μH with $D = 6\text{in}$ gave a Q of 650.
 - A 3mH variometer rated at 2 amps using close wound 0.33mm Litz (DC resistance 25 ohms) on a former of $D = 3\text{in}$ gave Q of 60.
 - A 2mH variometer on a former of $D = 3\text{in}$ the using 0.4mm solid wire ran hot at 2 amps and had a Q of 40.
- Conductor current (I) in the worked example is 1.45mm x



$$5 = 7.25 \text{ amps.}$$

Dissipation with key down = $7.25 \times 7.25 \times 6 = 315$ watts!
Obviously Litz wire is justified.

Fig 4.22: Detail of inner coil bearings.

Voltage Rating

Another factor which may need to be considered is the voltage per turn, given by $I \times XL / n$ where XL is the reactance at minimum inductance and n is the active turns. The enamel insulation will be a limiting factor with close winding. 50 volts per turn is probably safe. If turns are spaced it is acceptable to use 1kV per mm of spacing for good quality formers. In the example the maximum antenna current is limited by 50 volts per turn to 2.8 amps at maximum inductance and about 7 amps at minimum inductance with close winding. Somewhat thinner bare wire with a small amount of space requiring grooved formers will allow greater current with possibly higher Q.

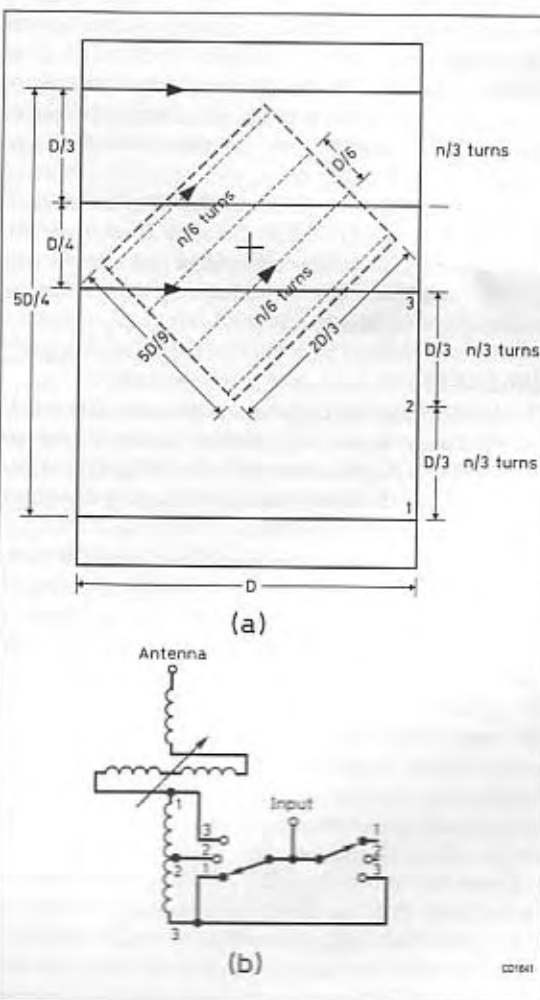


Fig 4.23: (a) Normalised variometer design. (b) Circuit diagram of the variometer.

5

Specialist Techniques

Very Narrow Bandwidth Techniques

By Rik Strobbe ON7YD

There are several reasons why the signal to noise ratio (SNR) of DX amateur radio signals on 136kHz is often very low:

- A wavelength of about 2.2km (1.35 miles) makes any amateur antenna small and inefficient.
- 1 watt ERP power limit.
- High noise levels.
- Strong commercial stations (100kW and more) very close to the 136kHz allocation.

One way of improving the SNR is to reduce the receiver bandwidth and thus have fewer unwanted signals and less noise whilst leaving the level of the wanted signal unchanged. But the reception of any signal requires a minimal receiver bandwidth depending on the type of modulation. SSB has a typical bandwidth of 2.4kHz while the bandwidth of a CW signal is dependent on the speed, but in any (practical) case is less than 100Hz. The use of a filter with a narrower bandwidth than that of the transmitted signal will distort the signal.

Here I will describe a technique that allows communication with signals far below the noise level. It can be used to make a basic QSO and I think that in some way the 'value' of such QSOs can be compared to Meteor Scatter QSOs on VHF.

Bandwidth

The dominant and most efficient mode on 136kHz is CW. As mentioned before the minimal bandwidth at the receiver side is determined by spectrum of the transmitted signal. In case of CW it is the (keying) speed that puts a limit to the minimal bandwidth.

An accepted method to measure the speed of CW is the PARIS system. The word Paris has a lengths of exactly 50 'dots', word spacing included. Based on this system a CW signal of 12 words per minute (WPM) means 600 'dot lengths' per minute or 10 'dot lengths' per second. But as each dot is separated by space of the same length, the actual length of a 'dot' is the double. If a continuous series of dots is given at 12WPM the result is a 5Hz square wave. If an RF signal is keyed with this series of dots you will get a carrier and 2 sidebands at 5Hz, resulting in a 10Hz wide signal.

Depending on how 'hard' the keying is, more sidebands further away from the carrier will be created but these do not contain additional information and can be considered as a waste of energy (and a source of interference to others). So basically the minimum bandwidth that is required

to receive an undistorted CW signal is:

$$B = 0.833 \times \text{WPM (Hz)}$$

Assuming that the only noise source is a frequency independent (white) noise, the total receiver noise will be directly proportional to the receiver bandwidth. Taking a 12WPM CW signal as a reference and assuming that the receiver bandwidth is optimized to the transmission speed the table below shows the signal-to-noise ratio (SNR) improvement that can be achieved by reducing the CW speed:

Speed	Optimum bandwidth	SNR vs 12WPM
12WPM	10Hz	0dB
8WPM	6.67Hz	+1.8dB
4WPM	3.33Hz	+4.8dB
1 sec / dot	1Hz	+10dB
3 sec / dot	0.33Hz	+14.8dB
10 sec / dot	0.1Hz	+20dB

It is clear that a significant SNR improvement can be achieved by reducing the CW speed. On 136kHz a dot length of 3 seconds has become a kind of standard. At these very slow CW speeds it becomes rather difficult to copy the signal by ear as you would almost need a chronometer to time the dots and dashes. Furthermore the frequency of the signal needs to be very stable as smaller bandwidths are used. Fortunately this not a big problem on 136kHz where a frequency stability of 0.1Hz is not difficult to achieve.

Another problem is that filters become more and more complicated to build as the bandwidth becomes smaller. Furthermore, tuning the band for a signal is rather laborious at bandwidths below 1Hz.

One way of overcoming many of these problems is the use of Digital Signal Processing (DSP).

Digital Signal Processing (DSP)

DSP basics

Digital Signal Processing is one of these rather technical expressions that seem to be the domain of specialised electronics engineers and some 'techy' hams. Until very recently special (and rather expensive) hardware was needed to perform DSP. But now a Pentium PC with soundcard can replace all the special hardware, and the software you need is available for free as is described later. As the expression Digital Signal Processing says, the analog (input) signal is converted to digital, then

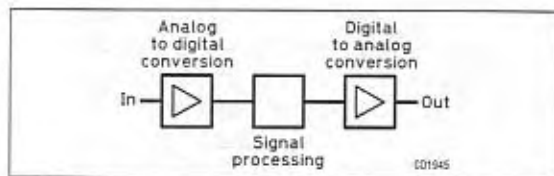


Fig 5.1: Simplified diagram of a digital processing system.

processed and eventually converted back to an analog (output) signal as shown in Fig 5.1.

The conversion of the analog signal to a digital form is done by analog-to-digital conversion (ADC). The most basic version of ADC is often done when we use a voltmeter to determine the value of a voltage. With DSP this 'reading of voltages' is done automatically at a known time interval – see Fig 5.2 – and this is called sampling. The result is a series of measurements, where we know the measured voltage and the time when it was measured.

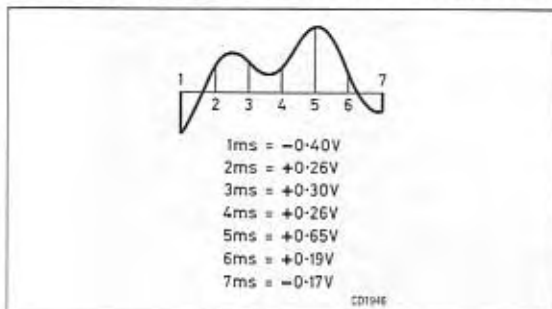


Fig 5.2: Example of digital sampling of an analogue signal, with samples being taken every milli-second.

These data are processed digitally. In practice that means that they undergo a series of calculations. The result can be interpreted as digital data or eventually converted back to an analog signal as shown in Fig 5.1. All manner of processes can be performed on a signal using DSP, such as filtering, reducing bandwidth, time multiplexing of several signals *etc*; but here we will only discuss the filtering of a signal.

Fast Fourier Transform (FFT)

Although there are several methods of filtering a signal digitally the most common is by the use of the Fast Fourier Transform (FFT). The basic idea behind this transform is that any signal can be seen as the sum of a series of sinusoidal signals, where each sine can have a different amplitude and phase as shown in Fig 5.3.

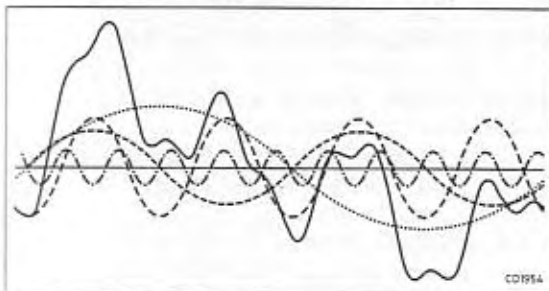


Fig 5.3: The complex (solid line) signal is equal to the sum of the three dotted sinewaves.

Because there is a lot of calculating involved Fourier transformations take a lot of 'computing time'. To reduce this a special algorithm was developed to enhance the speed of the Fourier transformations; this algorithm is called the Fast Fourier Transform (FFT).

When Fourier transform is performed on a signal it is

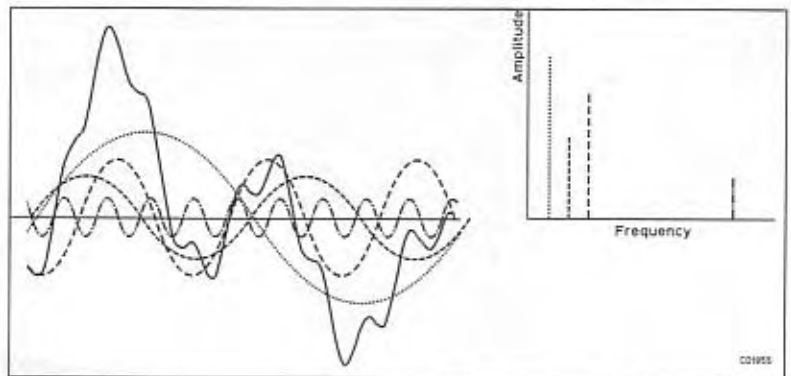


Fig 5.4: On the left is an example of the sinewaves that make up a complex signal shown in the time domain. A frequency domain display of these sine waves is shown on the right.

broken up into a number of sinewaves and the amplitude and phase of these sinewaves is then calculated. Each of these sinewaves represent a certain frequency (or frequency band [channel?]) and from sum of sines (and their amplitudes and phases) we can reconstruct the frequency spectrum of the measured signal.

The 'quality' of the reconstructed frequency spectrum depends on:

- the sample rate (interval between two A-D conversions)
- the sampling time for one transform
- the number of bits of the A-D converter

The sample rate determines the maximum frequency of the spectrum: the maximum frequency that can be reconstructed is 50% of the sampling frequency.

eg: If we take a sample every 0.2ms (equals a sampling frequency of 5kHz) the maximum frequency that can be reconstructed is 2.5kHz.

The sampling time determines the frequency accuracy (or the bandwidth of each 'channel'); the frequency accuracy is equal to one over the sampling time.

eg: If we take a sampling time of 0.1 seconds the frequency accuracy (or channel bandwidth) will be 10Hz. This means that in the series of sinewaves of the Fourier transform each sinewave will represent a 10Hz wide channel.

The number of samples in a Fourier transform has to be a power of 2 (2, 4, 8, 16,256,65536, ...). Although you can take any number of samples and just add a series of 'zeros' until you get a power of 2 it is more practical to choose the correct ratio between sample rate and sampling time in order to get the right number of samples.

eg: if we have a sample rate of 0.2ms we will not take a sampling time of 0.1 seconds, that would result in 500 samples, but a sample time of 0.1024 seconds in order to get 512 samples ($= 2^9$). The result of the Fourier transform will be a series of 256 sines where each sine represents a 9.766Hz wide channel between 0Hz and 2.5kHz.

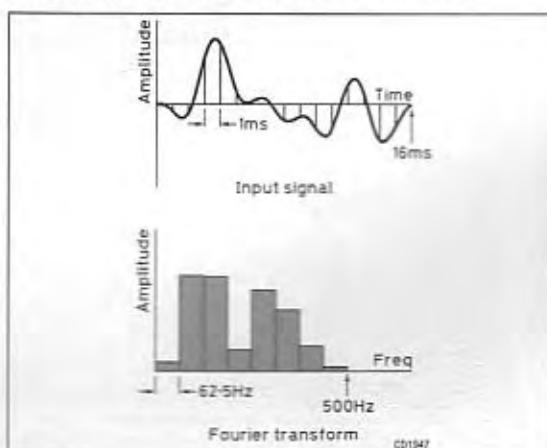


Fig 5.5: the Fourier transform of 16 samples at a rate of 1ms results in a series of 8 sines that each represent a 62.5Hz wide channel between 0 and 500Hz.

A further example is shown in Fig 5.5. A Fourier transform of 16 samples at a rate of 1ms results in a series of 8 sines that each represent a 62.5Hz wide channel between 0 and 500Hz:

The number of bits of the A-D converter determines the dynamic range of the spectrum. In practice (using the PC soundcard) we can choose between an 8-bit or 16-bit A-D conversion.

eg: For a 8-bit A-D conversion we have $2^8 = 256$ levels and the dynamic range will be $20 \cdot \text{Log}(256) = 48\text{dB}$. For a 16-bit A-D conversion we have $2^{16} = 65536$ levels and the dynamic range will be $20 \cdot \text{Log}(65536) = 96\text{dB}$.

Very narrowband modes
QRSS

QRSS is extreme slow speed CW. The name is derived from the Q-code QRS (reduce your speed). To take advantage of the very narrow bandwidth of the transmitted signal an appropriate filter at the receiver end is needed. Making a 'software filter' using FFT has some advantages over the hardware filter. One of the main advantages of using a software filter for the reception of slow CW signals is that FFT does not give one single filter, but a series of filters, which can be used to monitor a complete spectrum at once. This means that you do not have to tune to the exact frequency of the signal, which would have to be done very precisely at sub-Hertz bandwidths. It is also possible to monitor more than one QRSS signal at the same time.

The long duration of the dots and dashes and the ability to monitor more than one channel at the same time makes aural monitoring, using this technique, impractical.

The solution is to show the results of FFT processing on screen rather than making it audible. The result is a graphic where one axis represents time, the other frequency and the colour (or intensity) represents the signal strength. If the vertical axis represents time we call it a waterfall display, while it is called a curtain display if the horizontal axis represents time. All this may sound complicated but it is easy to understand when you see an example in Fig 5.6 (curtain display):

In April 2000 Geri Kinzel (DK8KW) did some measurements to compare QRSS with normal (aural) CW. He used a calibrated frequency synthesizer (Adret 2230), a 0-120 dB attenuator in 1dB steps. (Schlumberger BMD500) and a Praecitronic MV61 Selective Level Meter.

Using a BNC T-connector, normal band noise (including LORAN lines) on 137.500kHz ($\pm 50\text{Hz}$) was mixed with the output of the frequency synthesizer. The attenuator was set so that a 0dBm (50Ω) signal from the synthesizer corresponds to a -80dBu (0dBu = 0.775V into 75Ω = +9dBm, -80dBu = -71dBm) signal at the MV62 ($\pm 1\text{dB}$).

The band was quiet, with a background noise around -110dBu (S4, -101dBm) and LORAN lines clearly visible. Using the 100Hz bandwidth of the MV62 and the

cascaded 250Hz / 500Hz CW filters of the IC-746 the signal was monitored aurally, while at the same time viewing Spectrogram software. The parameters were: QRSS dot length 3-5 seconds, 5.5k sample rate, 16-bit mono, 16384 points FFT = 0.3 Hz resolution, 60 dB scale, 300 ms time scale and 10 x average. The results were as follows:

Signal strength at RX input	Comment
-100dBu/-91dBm	good audible CW (S6)
-110dBu/-101dBm	CW signal equal to noise level (S4), can just be copied
-115dBu/-106dBm	boundary for aural CW, signal just detectable by ear
-125dBu/-116dBm	perfect readable QRSS signal ('O' report)
-130dBu/-121dBm	good readable QRSS signal ('M' report)
-135dBu/-126dBm	just detectable QRSS signal ('T' report)
-140dBu/-131dBm	signal not detectable

QRSS would appear to have a 20dB signal level advantage over normal (aural CW). This means that the minimum detectable and/or readable QRSS signal for communication is 20dB below the signal that can just be detected and/or decoded by a trained CW-operator's ear (assuming a CW-operator's ear / brain bandwidth is 30Hz).

At a speed of 3 seconds per dot a very basic QSO will take about 30 minutes. Changing QRN levels and/or propagation during this period can, understandably, have a detrimental effect on a QSO. A new transmission mode has been developed that enhances the average speed by a factor of 2.5 to 3.

Dual Frequency CW (DFCW)

When the nature of CW is first analysed it appears to be a digital mode where 'key down' represents a logic '1' and 'key up' a logic '0'. But another approach it is to see it as a mode with 3 'logical states':

- the 'dash' (3 periods of key down + 1 period of key up or '1110')
- the 'dot' (1 period of key down + 1 period of key up or '10')
- the 'character space' (2 periods of key up or '00').

The spacing between words is 3 character spaces. So there are two elements that play a role: the presence/absence of a signal and the duration of the signal. As CW was intended to be received by ear the different duration of the signals is essential, but it lengthens the time needed to transmit a text.

In Dual Frequency CW (DFCW) the element 'duration' is replaced by the element 'frequency'. So dots and dashes no longer have a different length but they are transmitted on a different frequency. Because of this frequency shift there is no 'space' required between the dots and dashes and the character space can be reduced to the same (dot) length.

When the idea of DFCW first was introduced there was a lot of scepticism about the readability of frequency shifted signals, but in practice it seems rather easy to read it from the screen. To improve readability, especially during a sequence of dots or dashes, a short space (typically one third of a dot length) is added between the dots and dashes. This reduces the average speed slightly, but



Fig 5.6: Curtain display showing HB9ASB calling CQ.

improves the readability and reduces the duty cycle, which is less demanding for the PA.

At a speed of 3 seconds per dot a CQ that takes 5 minutes and 30 seconds in QRSS will only take 1 minute and 54 seconds in DFCW, as shown in Fig 5.7. The speed advantage of DFCW over QRSS can be used in 2 ways, either by reducing the duration of a QSO or by increasing the dot length and working at a narrower bandwidth. This means that, for the same duration of a QSO, the dot length in DFCW can be increased by 2.5 to 3, resulting in a 4 to 5dB SNR improvement.



Fig 5.7: The example of 'CQ ON7YD K' in QRSS and DFCW, being sent at the same speed.

Future developments of extreme narrowband modes

Over the past year a dot length of 3 seconds has become a kind of informal standard for QRSS and DFCW, as practical tests have shown best results for this speed. Most amateurs use the Spectrogram software with a sample rate of 11kHz and blocks of 16384 samples, which gives a sample time of about 1.5 seconds. At first sight it does not appear obvious why the sample time is only half the dot length; would it not be better to either make the sample time longer or the dot length shorter?

However, there is a good reason why the sample time is so much shorter than the dot length and that is because the transmitter and receiver are not 'synchronized'. This means that a sample block (generated at the receive end) can begin somewhere in the middle of a dot (generated at the transmitter end) and vice versa. The effect of this lack of synchronization on the processed signal is shown in Fig 5.8.

To ensure that at least one sample block falls completely within each dot (or dash, space) a sample block can not be longer than half the dot length.

If some kind of 'synchronisation' can be created between Tx and Rx it would be possible to (almost) double the sample block duration without increasing the dot

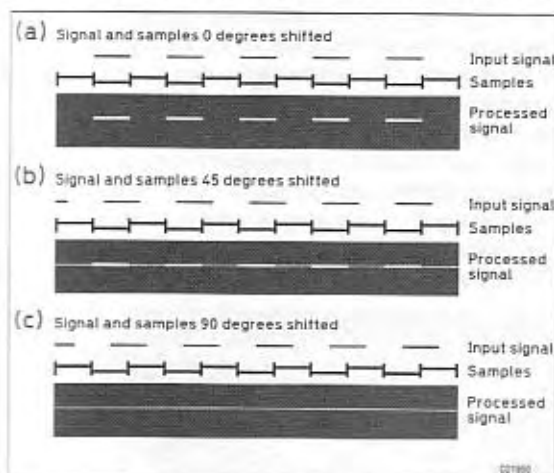


Fig 5.8: When the dot length and sample block have the same duration the signal is received clearly provided (a) the signal and sample time are in phase, ie synchronised. If the signal and sample time are out of phase by (b) 45° the signal becomes 'smeared'. If the phase slips as much as 90° then the signal is lost.

length and thus achieve a 3dB gain. When DFCW is used and the dot length is increased to 10 seconds a QSO will take about the same time as a 3 sec/dot QRSS QSO. The transmitter and receiver 'synchronise' problem could be overcome with appropriate software. This would give a 1 second accuracy to a sample block of 8 seconds (with a 2 seconds interleave) using a dot length of 10 seconds, as shown in Fig 5.9.

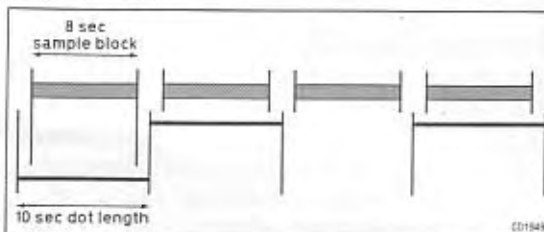


Fig 5.9: Suggested method of synchronizing the receiver to the transmitter with slow CW.

Compared with the 'traditional' QRSS a gain of over 7dB can be achieved while the duration of the QSO is about the same. Timing errors (between Tx and Rx) up to 1 second will not affect the SNR.

The programmers of Spectran have used an alternative solution. Instead of using a complete new set of samples for each FFT they take only part new data and 'shift' the existing data up in the data-array used to perform the FFT as shown in Fig 5.10.

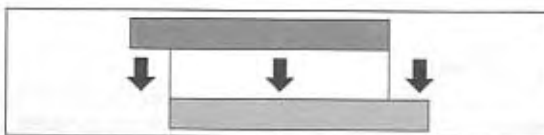


Fig 5.10: Data shift method used in Spectran. Assume that 4096 data points are used to perform a FFT. Instead of using a complete new set of data for the next FFT the 128 'oldest' points are removed (in position 3969 to 4096 in the data-array). The data points in position 1 to 3968 are then shifted upward to the end of the data-array. Positions 1 to 128 are then filled with new data. This procedure is repeated for every FFT.

The data shift method has the advantage that the 'duration' of the FFT array can be almost as long as the duration of a dot, but there are also some disadvantages. First of all the workload for the computer increases significantly; in case of the above example the computer has to perform 32 FFT's in the time that with the 'traditional' method only 1 FFT calculation is needed. On screen this method also causes some 'blur' at the beginning and end of the dots as shown in Fig 5.11.

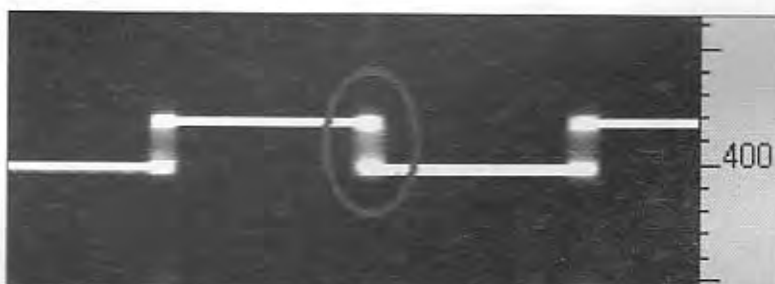


Fig 5.11: Blurring at the end of the dots in a data signal caused by the computer workload when using the data shift method of FFT processing.

Receiving and Transmitting Slow CW

By Peter Dodd G3LDO

This is a description of how to receive and transmit Slow CW or QRSS using a standard PC installation.

Receiving Slow CW

The basic theory has already been described by ON7YD. What follows are practical considerations and some examples of what can be achieved using the technique.

There are two popular programs suitable for producing these displays currently in use at the time of writing: Spectrogram and Spectran.

Spectrogram was written by Richard Home, whose main interest is recording and analyzing bird song. This program provides various methods of saving files for subsequent analysis.

Spectran was written by I2PHD and IK2CZL specifically for amateur radio for receiving weak signals although, at the time of writing, is still in the beta stage.

These programs use the A/D circuits of a soundcard to convert the analogue signals into a digital form for processing and use the rest of the computer to process the data and display it on the screen. The display shows the signal as frequency (normally on the vertical axis) plotted against time (normally on the horizontal axis). Signal strength is shown as either colour or intensity. This display is often known as a waterfall display.

These programs can be downloaded from the Internet; details of how this is done are given in the Appendix. Once your selected program is installed on your computer all you have to do is run it up. The output from the receiver, eg the headphones jack, is connected via a screened lead to the input of the soundcard. It's as simple as that.

There are two inputs to the soundcard – a line input and a microphone input. Use the line input socket, as the microphone input is too sensitive for this application. More information on the soundcards can be found in the Appendix.

Both programs have a frequency scale or pointer. For the frequency scale readings to be meaningful it is important that the receiver is set up as follows:

By general consensus most Morse slow CW QSOs are carried out at the top end of the band, above 137kHz. If

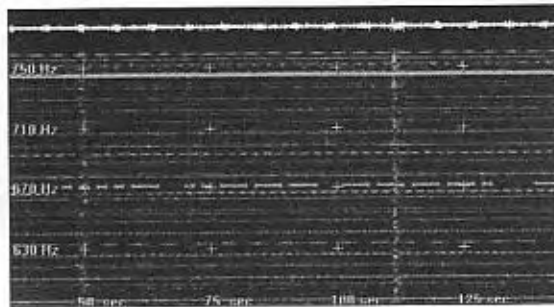


Fig 5.12: Signal received using Spectrogram from IK1ODO in a 136.670kHz slow CW QSO on 28 Feb 1999. Conditions are good and I am also given a RST 429 report indicating the signal, as well as being seen, was just audible. The masses of horizontal lines are sidebands from the 100kHz Loran station at Lessay in northern France. The display bandwidth is around 160Hz.

the receiver is set exactly on 137kHz SSB upper sideband the signal shown in the Spectrogram display in Fig 5.12, on 670Hz, means that the received signal is on 137.670kHz. On Spectrogram the time and frequency grid shown can only be displayed when a recorded signal is played back for analysis; however, during real time monitoring or recording a cursor can be placed over the signal and the frequency displayed in a box at the bottom of the display.

With Spectran the frequency is displayed more directly on a scale. Because the display can be shown vertically or horizontally the frequency scales are shown on both axis and time is not scaled.

An example of a Spectran display is shown in Fig 5.13.

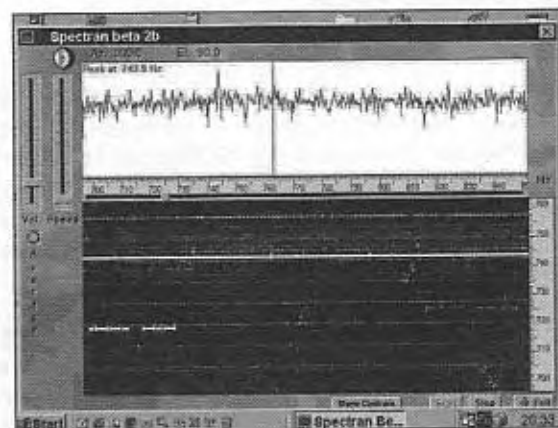


Fig 5.13: Spectran display, with a strong signal from G3YXM. A weak signal can be seen at 137.780kHz, which was at first thought to be some sort of delayed echo from G3YXM. It turned out to be the 20mW station, I5TGC, calling CQ. Later I5TGC QSYed to 137.725kHz and called CQ again. Note the carrier on 137.4kHz, which also peaks above the noise in the spectrum analyzer display.

The main difference is that the display has a better resolution and an additional frequency/amplitude (spectrum analyzer) display is included in the top of the display. There is also a spectrum analyzer display with Spectrogram, which is calibrated so that it can be used as a spectrum analyzer. It cannot be run at the same time as the waterfall display.

The Spectran software is under development and most examples shown here are using earlier versions. Not shown in Fig 5.13 is a control panel display, which can be accessed without switching off the main display. The latest version of Spectran, at the time of writing is shown in

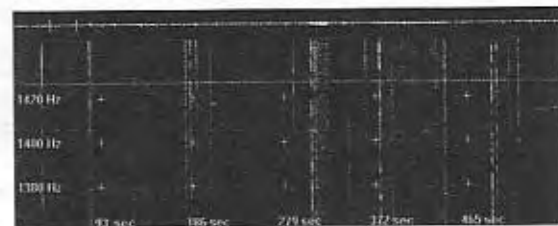


Fig 5.14: 71.8kHz signals from G3LDO, received by CT1DRP in Porto using spectrogram. It shows the 71kHz side of a crossband QSO with I5TGC. I am giving an 'M' report. The vertical lines indicate heavy QRN.

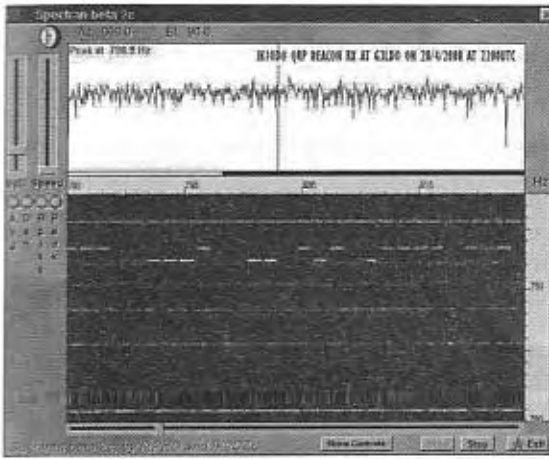


Fig 5.15: DFCW signal received from IK1ODO's QRP beacon. The bandwidth of the display is 18Hz so that the DFCW shift of 1Hz is clearly seen. The dots are on the lower line and the dashes on the top line.



Fig 5.16: Signals from G3YXM (top) and G3LDO (bottom) received by OH1TN. This displays the ability of the mode to monitor several signals at the same time.

Fig 5.19, being used to decode a Hellschreiber with a bandwidth of only 2.5Hz. Fig 5.19 also shows the control panel with selectable parameters displayed.

The examples shown in Figs 5.14, 5.15 and 5.16 show the potential of this type of software.

Transmitting Slow CW

The simplest way of transmitting slow CW is to use a memory keyer with a modified speed control. This can simply be a resistor in series with the speed control. However, this is not a very convenient method from an operating point of view. A much better way is to use a computer with a very simple circuit on the output of the COM serial port, as shown in Fig 5.17.

Then you can see what has been entered, which is very useful when composing a response to a message unfolding on the slow CW display. There are two such programs available, SLOWCW, and QRS; both are available from the RSGB website (see Appendix).

Both programs require that a serial port (COM) is assigned before you can drive a transmitter. Be sure *not* to assign the same port already assigned to the mouse.

SLOWCW was written by Andy Talbot, G4JNT and it appears to be able to run on any IBM computer. I use a very old Toshiba laptop with no hard drive and just one floppy drive (obtained from a computer junk shop for £5) as a dedicated slow speed keyer.

The text for the main CW message is then entered when prompted. All parameters will then be written to a file named CWSLOW.INF, which can be edited manually. When the program is subsequently run, these parameters are then offered as defaults. Alternatively the program may be started immediately by adding /I to the command line ie CWSLOW /I. This is useful for restoring after power cuts.

In some cases the standard spacing does not give maximum readability to the signal during DX contacts. The spacing can be modified with this program as a percentage; I generally use 20% increase in spacing for DX slow CW contacts.

This program provides transmitter control as well as

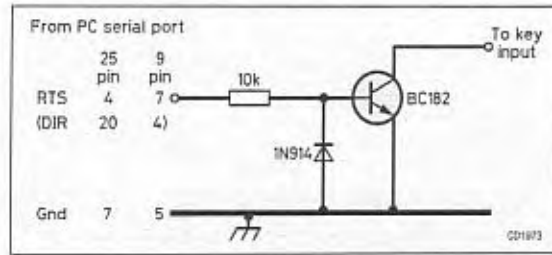


Fig 5.17: Slow CW interface for a PC serial port. Note the pin numbers for 9-way D-type connector and the 25-way D-type connector. (a) CW keying line uses the Request to Send line - RTS. (b) Transmitter control is on the Data Terminal Ready line (DTR).

keying. This means that the transmitter can be switched on and off at preselected times.

QRS is written in Delphi 1.0 under Win3.11 by Rik Strobbe, ON7YD, but will work also under Win95/98. It also requires TCommPortDriver. See Appendix.

The features of QRS are as follows:

- slow CW at dot lengths between 0.1 and 60 seconds.
- supports QRSS and DFCW modes.
- fast CW up to 20WPM (but reduced accuracy above 8WPM due to the multitask functionality).
- PTT (or FSK) and keying via a serial port using a very simple interface (COM1 to COM16).
- FSK tones also available via internal speaker for modulating a SSB transmitter.
- 10 text buffers each of 70 characters available.
- repeat function allows any text to be repeated (once to endless) at a selectable interval time.
- transmission can start immediately or at any given time.
- functional under multitask environment (eg can be used simultaneously with receiving software).
- available characters are A-Z, 0-9 ? / AR SK BK CT (@ = AR \$ = SK. % BK. * = CT).
- test function for key and PTT/FSK.
- 'QSO mode' for optimum screen sharing with Spectrogram.
- 'beacon mode' enables a number of dashes to be sent at the end of the text.
- QSK facilities to allow listen through during a QRSS transmission.
- selectable dash/dot ratio allows an average 20% time saving in QRSS mode.
- alarm function available for acoustic warning when the transmission is about to end.
- CW sidetone (400Hz or 800Hz) available via internal PC speaker.
- QRSS/DFCW transmissions can be started and ended with a 6 or 12 WPM CW identification.

Operating Practice

The following are notes from ON7YD on guidance on operating procedures for QRSS and DFCW. LF operators have now accepted these procedures.

The (unofficial) QRSS/DFCW segment is 137,600Hz - 137,800Hz. Most activity is between 137,700Hz and 137,750Hz

You need a stable Tx with a long-term stability of 5Hz as a bare minimum, although 1Hz or better is recommended.

Keep your CQs short: eg. CQ G3XXX K not CQ CQ CQ DE G3XXX G3XXX G3XXX PSE K

The report system is the TMO system (similar to that used in EME):

- T = signal traces seen but not good enough for a QSO
 - M = weak signal but good enough for a QSO
 - O = perfect copy
- For QRSS a dot length of 3 seconds is recommended,

although 5 seconds dots can be used if very weak signals are anticipated. For DFCW a shift of 5Hz (with 'dash' being the higher frequency) and a key gap of 1 second is recommended

During a QSO (once you are sure that both stations have the calls OK) you can use the suffixes instead of complete calls.

If you see a QSO coming to an end and you want to contact one of the stations you can start calling this station while the previous QSO is going on, just call on another frequency.

When replying to a CQ it is recommended that you do *not* do this exactly on the frequency of the other station, this will avoid QRM in the event of more than one station is responding.

A basic QRSS (or DFCW) QSO could look like this:

- CQ ON7YD K
- ON7YD G3XDV K
- G3XDV YD OOO K
- YD XDV OOO K
- XDV YD TU 73 K
- YD XDV GL 73 SK

Low Speed Hellschreiber on LF

By Andy Talbot G4JNT

The Hellschreiber Mode

Hellschreiber is a mode developed in Germany in the 1940s and literally translates as "Light Writer". It is a means of sending text characters over a radio link in a manner similar to fax. The original scheme used on-off keying of a carrier to create light or dark pixels on photosensitive paper on a rotating drum using a scanning system, similar to television, for each letter. Owing to the need to send one pixel at a time the letters must necessarily slant and various fonts were developed to take care of this and improve the appearance.

With the arrival of computers a simpler method of generating and displaying the signals was possible and the method went through a re-birth amongst a few amateurs with custom software to recreate the raster timing information. Other systems than on-off keying were tried and a very good scheme used frequency steps for vertical lines, and timing for the horizontal lines. By transmitting the 'vertical' frequencies simultaneously, true vertical lines were now possible. Obviously the transmitted signal is more complex than the on-off keyed version, as anything up to seven frequencies may have to be transmitted at the same time. Spectrogram type software, where a colour 'waterfall' display shows frequency by vertical position, time along the horizontal axis and amplitude by colour or brightness allows the text to appear directly on the screen once frequency span and timing have been appropriately adjusted. This became known as Multi Tone Hellschreiber or MT Hell.

To save the need for linear transmitters and complex signal generation an intermediate mode, called Sequential Multi Tone Hell – SMT Hell – was developed. This still

uses frequency for the vertical information but each pixel is now sent sequentially. Letters lean as vertical lines are no longer possible, but are quite readable when viewed on a spectrogram and only one frequency needs to be generated at a time allowing non-linear transmitters. The signal is now an on-off keyed tone taking on any one of eight or more discrete frequencies.

All Hellschreiber modes are catered for in comprehensive software; see Appendix. Versions for the 56002EVM have also been developed by G3PLX and others. All these are for normal typing speed, and the tones are spaced over a bandwidth of several hundred Hz to 1kHz. Versions using the Windows operating system allow any font available to the operating system to be employed, but custom Hellschreiber fonts usually appear to work best, especially for the sequential modes.

Direct Signal Generation

After developing an AD9850 Direct Digital Synthesizer design using a PC to directly control the frequency being generated, I took this SMT Hell a stage further using the DDS to generate each tone. Designed for LF without having to use a transverter or SSB transmitter, the DDS generates the RF carrier directly and, using a 5MHz clock, any frequency can be generated to a resolution of 0.0011Hz. Frequency spacing was kept low and suitable timing pulses for the horizontal pixels were chosen experimentally based on the software likely to be used for decoding the signals. Software was written to take a simple raster font – an 8 x 8 dot one stored in the operating system and used for the VGA 50 line mode in DOS, then translate this into the codes needed to drive the DDS board, sent at the correct timing intervals.

So far on air tests have been carried out using a 10Hz top to bottom span with a pixel duration of 0.6 seconds. For an 8 x 8 font this therefore means a character takes 38.4 seconds to send. As I write this a 5Hz wide, 1.2 second per pixel transmission is going out taking 76.8 seconds per character. Other fonts were tried, in particular a 7 x 5 one which is probably the smallest giving full alphabetic character resolution and similar to the original Hellschreiber font. This was reported as not being so readable. Others, (used for the DOS 25-line and 30-line modes) having a higher pixel resolution are also available but have not yet been

Fig 5.18: Hellschreiber signal transmitted by G4JNT on 137.6kHz using a 10Hz bandwidth and received by G3LDO (58km) using Spectrogram with a TS-850. The receiver was set to USB and 12dB of attenuation added for good measure.



tried on air, taking longer to send per character.

If this software were ported to the Windows operating system, the whole set of fonts available there could be used, but experience with the conventional Hellschreiber software shows no advantage in using these over the simple dot raster ones.

Receiving the Signal

Hellschreiber signals can be received using Spectrogram or Spectran, described earlier. Wider bandwidth signals can be decoded using Spectrogram; see Fig 5.18. For the 10Hz wide version a frequency resolution of either 0.125Hz or 0.25Hz is suitable, giving a vertical display of around 30Hz and means the letters appear over about a third of the screen height. Scrolling speed has to be adjusted to give the correct spacing; too fast and the letters break up into individual sloping lines; too slow and the characters appear too thin and unreadable. A speed setting of 20-40% usually gives acceptable results.

Narrower bandwidths can be used and Spectran is probably the best readily available software for this purpose. An example of a decoded 2.5Hz signal is shown in Fig 5.19. This is the latest version of Spectran and the control panel is switched in showing selectable prime-terms.

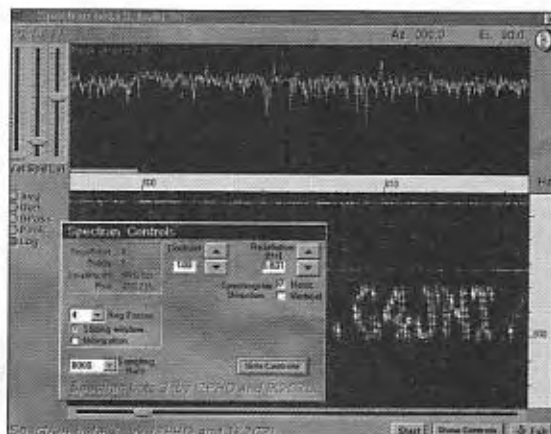


Fig 5.19: Hellschreiber signal transmitted by G4JNT on 137.6kHz using a 2.5Hz bandwidth and received by G3LDO (58km) using Spectran with a TS-850. Here the total bandwidth displayed on the screen is 9Hz and the signal occupies only 2.5Hz bandwidth. Note that the control panel is switched in showing selectable prime-terms.

I also obtained very good results decoding using EVM-SPEC and EVM-DOP Dopplergram software for the 56002EVM.

For correct orientation of the letters, upper sideband needs to be used for reception resulting in the lowest carrier frequency appearing at the bottom of the trace. If LSB must be used, the screen has to be looked at through a top to bottom mirror to get the correct letter orientation! Alternatively, use vertical scrolling and view the screen sideways.

PSK31: A New Radio-Teletype Mode

by Peter Martinez, G3PLX

There is a gap opening up between the data transfer enthusiasts using the latest techniques and the two-way contact fans who are still using the traditional RTTY mode of the '60s, although of course using keyboard and screen rather than teleprinter. There is scope for applying the new techniques now available to bring RTTY into the 21st century.

This article discusses the specific needs of 'live QSO' operating, as opposed to just transferring chunks of error-free data, and describes the PSK31 mode which I have developed specifically for live contacts, which is now becoming popular using low-cost DSP kits, and which could become even cheaper as the art of using PC sound cards is developed by amateur radio enthusiasts.

What is Needed?

I believe that it is the error-correcting process used in modern data modes which make them unsuitable for live contacts. I have identified several factors; the first revolves around the fact that all error-correcting systems introduce a time-delay into the link. In the case of an ARQ link like AmTOR or PacTOR, there is a fixed transmission cycle of 450mS or 1.25sec or more, which will delay any key-press by as much as one cycle-period, and by more if there are errors. With forward-error-correction systems there is also an inevitable delay, because the information is spread out over a period of time. In a live two-way contact, the delay is doubled at the point where the transmission is handed over. I believe that these delays make such systems unpleasant to use in a two way conversation.

This is not so much a technical problem as a human

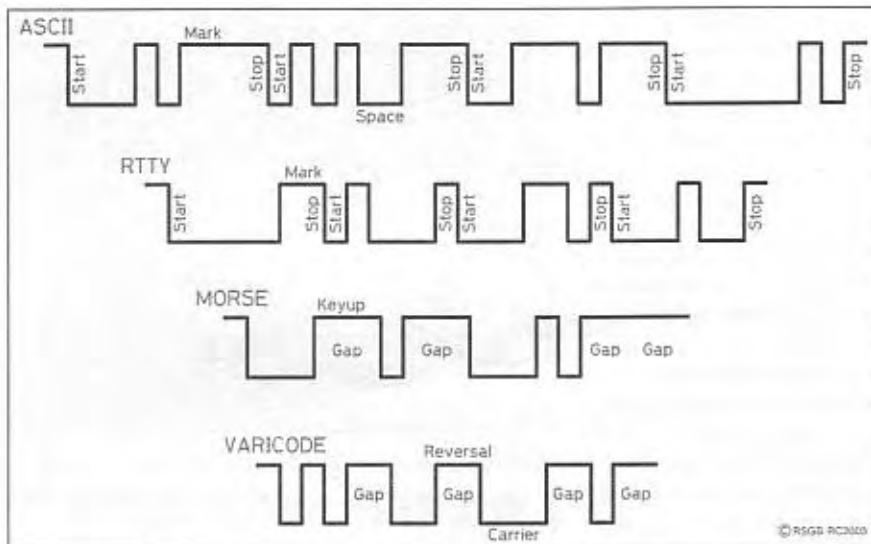
one. Another factor in this category is concerned with the way that the quality of the information content varies as the quality of the radio link varies. In an analogue transmission system such as SSB or CW, there is a linear relationship between the two.

The operators are aware of this all the time and take account of it subconsciously: they change the speed and tone of voice instinctively, and even choose the topic of conversation to suit the conditions.

In a digital mode the relationship between the signal-to-noise ratio on the air and the error-rate on the screen is not so smooth. The modern error-correcting digital modes are particularly bad at this, with copy being almost perfect while the SNR is above a certain level and stopping completely when the SNR drops below this level. The effect is of no consequence in an automatic mailbox forwarding link, but can badly inhibit the flow of a conversation. A third factor is a social one; with error-correcting modes you only get good copy when you are linked to one other station. The copy is decidedly worse when not linked, such as when calling CQ or listening to others. This makes it difficult 'getting to know' other people on the air, and there is a tendency to limit contacts to a few close friends or just mailboxes.

These factors lead me to suggest that there is a case for a transmission system that is not based on the use of error-correcting codes, when the specific application is that of live contacts. The continued popularity of traditional RTTY, using the start-stop system, is proof of this hypothesis: there is minimal delay (150mS), the flow of conversation is continuous, the error-rate is tolerable, and it is easy to listen-in and join-in.

Fig 5.20: Showing the word 'ten' keyed in ASCII, RTTY, Morse, and Varicode.



short codes for the more common letters, it is actually very efficient in terms of the average duration of a character. In addition, if we think of it in terms that we normally use for digital modes, Morse code is self-synchronising; we don't need to use a separate process to tell us where one character ends and the next begins. This means that Morse code doesn't suffer from the 'error-cascade' problem that results in the

Improving on RTTY

How, then, do we go about using modern techniques that were not available in the '60s, to improve on traditional RTTY? First of all, since we are talking about live contacts, there is no need to discuss any system that transmits text any faster than can be typed by hand.

Secondly, modern transceivers are far more stable in frequency than they were in the '60s, so we should be able to use much narrower bandwidths than in those days. Thirdly digital processors are much more powerful than the rotating cams and levers of the mechanical teleprinter, so we could use better coding. The drift-tolerant technique of frequency-shift keying, and the fixed-length five-unit start-stop code still used today for RTTY are a legacy of the limitations of technology 30 years ago. We can do better now.

PSK31 Alphabet

The method I have devised for using modern digital processing to improve on the start-stop code, without introducing extra delays due to the error-correcting or synchronisation processes, is based firmly on another tradition, namely that of Morse code. Because Morse uses

start-stop method getting badly out of step if a start or stop-bit is corrupted. This is because the pattern used to code a gap between two characters never occurs inside a character.

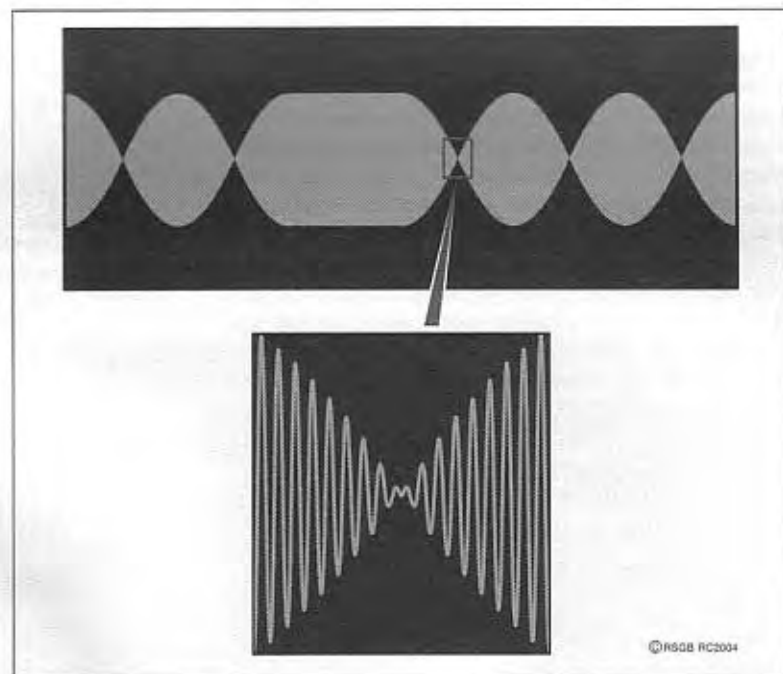
The code I have devised is therefore a logical extension of Morse code, using not just one-bit or three-bit code-elements (dots and dashes), but any length. The letter-gap can also be shortened to two bits. If we represent key-up by 0 and key-down by 1, then the shortest code is a single one by itself. The next is 11, then 101 and 111, then 1011, 1101, 1111, but not 1001 since we must not have two or more consecutive zeros inside a code. A few minutes with pencil and paper will generate more. We can do the 128-character ASCII set with 10 bits. I analysed lots of English language text to find out how common were each of the ASCII characters, then allocated the shorter codes to the more common characters. The result is shown in Table 1, and I call it the 'Varicode' alphabet. With English text, Varicode has an average code-length, including the '00' letter-gap, of 6.5 bits per character.

By simulating random bit errors and counting the number of corrupted characters, I find that Varicode is 50% better than start-stop code, thus verifying that its self-synchronising properties are working well.

The shortest code in Morse is the commonest letter 'e', but in Varicode the shortest code is allocated to the word-space. When idle, the transmitter sends a continuous string of zeros.

Fig 5.20 compares the coding of the same word in ASCII, RTTY, Morse, and Varicode.

Fig 5.21: Showing the waveform of BPSK sending the Varicode 'space' symbol.



PSK31 Modulation and Demodulation

To transmit varicode at a reasonable typing speed of about 50 words per minute needs a bit-rate of about 32 per sec. I have chosen 31.25, because it can be easily derived from the 8kHz sample-rate used in many DSP systems. In theory we only need a bandwidth of 31.25Hz to send this as binary data, and the frequency stability that this implies can be achieved with modern radio equipment on HF.

The method chosen was first used on the amateur bands, to my knowledge, by SP9VRC. Instead of frequency-shifting the carrier, which is wasteful of spectrum, or turning the carrier on and off, which is wasteful of transmitter power capability, the 'dots' of the code are signalled by reversing the polarity of the carrier. You can think of this as equivalent to transposing the wires to your antenna feeder.

This uses the transmitted signal more efficiently since we are comparing a positive signal before the reversal to a negative signal after it, rather than comparing the signal present in the dot to no-signal in the gap. But if we keyed the transmitter in this way at 31.25 baud, it would generate terrible key clicks, so we need to filter it.

If we take a string of dots in Morse code, and low-pass filter it to the theoretical minimum bandwidth, it will look the same as a carrier that is 100% amplitude-modulated by a sinewave at the dot-rate. The spectrum is a central carrier and two sidebands at 6dB down on either side.

A signal that is sending continuous reversals, filtered to the minimum bandwidth, is equivalent to a double sideband suppressed carrier emission, that is, to two tones either side of a suppressed carrier. The improvement in the performance of this polarity-reversal keying over on-off keying is thus equivalent to the textbook improvement in changing from amplitude modulation telephony with full carrier to double-sideband with suppressed carrier. I have called this technique 'polarity-reversal keying' so far, but everybody else calls it 'binary phase-shift keying', or BPSK. Fig 5.21 shows the envelope of BPSK modulation and the detail of the polarity reversal.

To generate BPSK in its simplest form we could convert our datastream to levels of $\pm 1V$ for example, take it through a low-pass filter, and feed it into a balanced modulator into which we also feed the desired carrier frequency. When sending continuous reversals, this looks like a 1V peak-to-peak sinewave going into a DSB modulator, so the output is a pure two-tone. In practice, we use a standard SSB transceiver and perform the modulation at audio frequency, or carry out the equivalent process in a DSP chip. We could signal a logic zero by continuous carrier and signal a logic one by a reversal, but I do it the other way round for reasons which will become clear shortly.

There are a variety of ways to demodulate BPSK, but they all start with a bandpass filter. For the speed chosen for PSK31, this filter can be as narrow as 31.25Hz in theory, but a 'brick-wall' filter of precisely this width would be costly, not only in monetary terms but in the delay time through the filter, and we are trying to avoid delays. A practical filter might be twice the baud-rate (62.5Hz) wide at the 50dB-down point, and have a delay-time of two bits (64mS).

For the demodulation itself, since BPSK is equivalent to double sideband, the textbook method for demodulating DSB can be used, but another way is to delay the signal by one bit-period and compare it with the direct signal in a phase comparator. The output is negative when the signal reverses polarity, and positive when it doesn't.

Although we could extract the information from the demodulated signal by measuring the lengths of the 'dots' and 'dashes', like we do by ear with Morse code, it will help to pick the data out of the noise if we know when to expect them. We can easily transmit the data at an accurately timed rate, so it should be possible to predict when to sample the demodulator output. This process is known as synchronous reception, although the term 'coherent' is some-times wrongly used. To synchronise the receiver to the transmitter, we can use the fact that a BPSK signal has an amplitude-modulation component. Although the modulation varies with the data pattern, there is always a pure tone component in it at the baud-rate, and this can be extracted using a narrow filter or a phase-lock loop, or the DSP equivalent, and fed to the decoder to sample the

demodulated data. Fig 5.22 shows block diagrams of a typical BPSK modulator and demodulator.

For the synchronisation to work we need to make sure there are no long gaps in the pattern of reversals. A completely steady carrier has no modulation, so we could never predict when the next reversal was due. Fortunately, Varicode is just what we need, provided we choose the logic levels so that zero corresponds to a reversal and one to a steady carrier. The idle signal of continuous zeros thus generates continuous reversals, giving us a strong 31.25Hz modulation. Even with continuous keying there will always be two reversals in the gaps between characters. The average number of reversals will therefore be more than two in every 6.5 bits, and there will never be more than 10 bits with no reversal at all. If we make sure that period, then the timing will pull into sync pretty quickly. By making the transmitter end a transmission with a 'tail' of unmodulated carrier, it is then possible to use the presence or absence of reversals to 'squell' the decoder so that the screen doesn't fill with noise when there is no signal.

Getting Going

So much for the philosophy and the theory, but how do you get on the air with this mode? At the moment, the route to getting on PSK31 is to obtain one of several DSP starter kits. These are printed circuit cards, usually with a serial interface to a PC, marketed by DSP processor manufacturers at low cost to help engineers and students become familiar with DSP programming. A number of radio amateurs have started to write software for these, not just for RTTY but also for SSTV, packet, satellite, and digital voice experiments. They have audio input and output and some general purpose digital input/output. The construction work needed is limited to wiring up cables, building a power supply, and putting the card into a screened box. The DSP software is freely available, as is the software that runs in the PC to interface to the keyboard and screen, and can be obtained most easily via the Internet. It would certainly be possible to construct a PSK31 modem in hardware, although I know of no-one having done this yet, but probably the most promising hardware platform for the future will be the PC sound card.

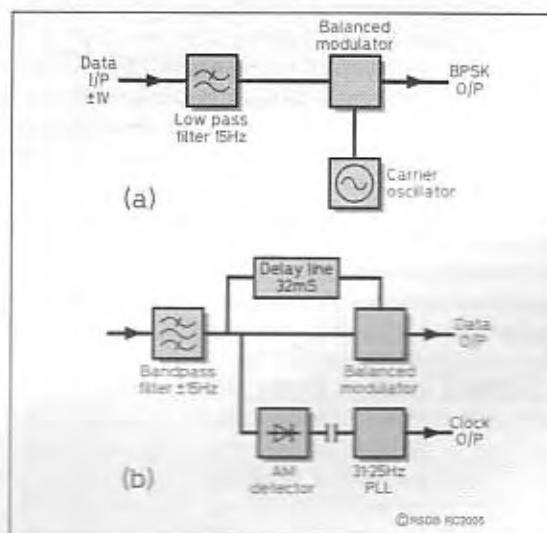


Fig 5.22: Block diagram of analogue BPSK modulator (a) and demodulator (b).

TABLE 1

The Varicode alphabet. The codes are transmitted left bit first, with '0' representing a phase reversal on BPSK and '1' representing a steady carrier. A minimum of two zeros is inserted between characters. Some implementations may not handle all the codes below 32.

ASCII	Varicode	ASCII	Varicode	ASCII	Varicode
0	1010101011	+	1110111111	V	110110101
1	1011011011	.	1110101	W	101011101
2	1011101101	-	110101	X	101110101
3	1101110111	.	1010111	Y	101111011
4	1011101011	/	110101111	Z	1010101101
5	1101011111	0	10110111	[111110111
6	1011101111	1	10111101	\	111101111
7	1011111101	2	11101101]	111111011
8	1011111111	3	11111111	^	1010111111
9	11101111	4	101110111	~	101101101
/feed	11101	5	101011011		1011011111
11	1101101111	6	101101011	a	1011
12	1011011101	7	110101101	b	1011111
c/ret	11111	8	110101011	c	101111
14	1101110101	9	110110111	d	101101
15	1110101011	:	11110101	e	11
16	1011110111	:	110111101	f	111101
17	1011110101	<	111101101	g	1011011
18	1110101101	=	1010101	h	101011
19	1110101111	>	111010111	i	1101
20	1101011011	?	1010101111	j	111101011
21	1101101011	@	1010111101	k	10111111
22	1101101101	A	1111101	l	11011
23	1101010111	B	11101011	m	111011
24	1101111011	C	10101101	n	1111
25	1101111101	D	10110101	o	111
26	1110110111	E	1110111	p	111111
27	1101010101	F	11011011	q	110111111
28	1101011101	G	11111101	r	10101
29	1110111011	H	101010101	s	10111
30	1011111011	I	1111111	t	101
31	1101111111	J	111111101	u	110111
space	1	K	101111101	v	1111011
!	111111111	L	11010111	w	1101011
"	101011111	M	10111011	x	11011111
#	111110101	N	11011101	y	1011101
\$	111011011	O	10101011	z	111010101
%	1011010101	P	11010101	{	1010110111
&	1010111011	Q	111011101		110111011
'	101111111	R	10101111	}	1010110101
(11111011	S	1101111	~	1011010111
)	11110111	T	1101101	127	1110110101
*	101101111	U	101010111		

PSK31 Operating

Since PSK31 performance is the same when calling, listening, or in contact, it's easy to progress from listening to others, to calling CQ, to two-way contacts and multi-way nets, but the narrow bandwidth and good weak-signal performance do mean learning a few new tricks. It's usual to leave the radio dial on one spot and fine-tune the audio frequency, listening through the narrow audio filter rather than the loudspeaker, and using an on-screen phase-shift display to centre the incoming signal within a few Hz. On transmit, since the envelope of the PSK31 signal is not constant (as is the case for FSK), it is important to keep the transmitter linear throughout. However, since the PSK31 idle is identical to a standard two-tone test signal, it is easy to set up. The worst distortion products will be at $\pm 45\text{Hz}$ at a typical level of 36dB below PEP.

A Second Look at Error Correction

After getting PSK31 going with BPSK modulation and the Varicode alphabet, several people urged me to add error correction to it in the belief that it would improve it still further. I resisted for the reasons that I gave above, namely that the delays in transmission, the discontinuous

traffic flow, and the inability to listen-in, all make error correction unattractive for live contacts. There is another reason. All error correcting systems works by adding redundant data bits. Suppose I devise an error correcting system that doubles the number of transmitted bits. If I wanted to keep the traffic throughput the same, I would need to double the bit rate. But with BPSK that means doubling the bandwidth, so I lose 3dB of signal-to-noise ratio and get more errors. The error correction system will have to work twice as hard just to break even! It is no longer obvious that error correction wins. It is interesting to note that with FSK, where the bandwidth is already much wider than the information content, you can double the bit-rate without doubling the bandwidth, and error correction does work. Computer simulation with BPSK in white noise shows that when the SNR is good, the error correction system does win, reducing the low error rate to very low levels, but at the SNR levels that are acceptable in live amateur contacts, it's better to transmit the raw data slowly in the narrowest bandwidth. It also takes up less band space of course!

However, there was the suggestion that error correction could give useful results for bursts of noise which cannot

be simulated on the bench, so I decided to try it and do some comparison tests. The automatic repeat (ARQ) method of correcting errors was ruled out, but forward error correction (FEC) which was added to modes such as AmTOR and PacTOR almost as an afterthought seemed to deserve a second look, provided the transmission delay was not too long.

I realised that comparing two systems with different bandwidths and speeds on the air would be difficult: adjacent channel interference would be different, as would the effects of multipath. There is, however, another way to double the information capacity of a BPSK channel without doubling its bandwidth and speed. By adding a second BPSK carrier at 90° at the transmitter and a second demodulator in the receiver, we can do the same trick that is used to transmit two colour-difference signals in PAL and NTSC television. I call this quadrature polarity reversal keying, but everybody else calls it quaternary phase shift keying or QPSK.

There is a 3dB signal-to-noise penalty with QPSK, because we have to split the transmitter power equally between the two channels. This is the same penalty as doubling the bandwidth, so we are no worse off. QPSK is therefore ideal for my planned comparison experiment: the adjacent channel interference, the SNR, and the multipath performance would be exactly the same for both.

In the next section I will think of QPSK not as two channels of binary data, but as a single-channel which can be switched to any of four 90° phase-shift values. By the way, the clock recovery idea used for BPSK works just as well on QPSK, because the envelope still has a modulation component at the bit-rate.

QPSK and the Convolutional Code

There is a vast amount of available knowledge about correcting errors in data which are organised in blocks of constant length such as ASCII codes, by transmitting longer blocks, but I know of nothing that covers error correction of variable length blocks like Varicode. However, there are ways of reducing errors in continuous streams of data which have no block structure, and this seems a natural choice for a radio link, since the errors don't have any block structure either. These are called convolutional codes, and one of the simplest forms does actually double the number of data bits and is therefore a natural choice for a QPSK channel carrying a variable length code.

The convolutional encoder generates one of the four phase shifts, not from each data bit to be sent, but from a sequence of them. This means that each bit is effectively spread out in time, inter-twined with earlier and later bits in a precise way. The more we spread it out, the better will be the ability of the code to correct bursts of noise, but we must not go too far or we will introduce too much transmission delay. I chose a time spread of 5 bits. The table that determines the phase shift for each pattern of 5 successive bits is given in the appendix. The logic behind this table will not be covered here.

In the receiver, a device called a Viterbi decoder is used. This is not so much a decoder as a whole family of encoders playing a guessing game. Each one makes a different 'guess' at what the last 5 transmitted data bits might have been. There are 32 different patterns of 5 bits and thus 32 encoders. At each step the phase shift value predicted by the bit-pattern-guess from each encoder is compared with the actual received phase shift value, and the 32 encoders are given 'marks out of ten' for accuracy. Just

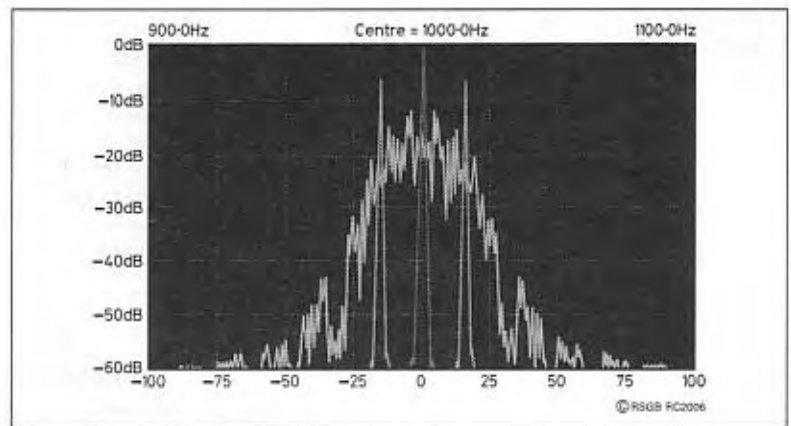


Fig 5.23: Showing the spectrum of the BPSK signal, idling and sending data, compared with an unmodulated carrier at the same signal level.

like in a knockout competition, the worst 16 are eliminated and the best 16 go on to the next round, taking their previous scores with them. Each surviving encoder then gives birth to two children, one guessing that the next transmitted bit will be a zero and the other guessing that it will be a one. They all do their encoding to guess what the next phase-shift will be, and are given marks out of ten again which are added on to their earlier scores. The worst 16 encoders are killed-off again and the cycle repeats.

It's a bit like Darwin's theory of evolution, and eventually all the descendants of the encoders that made the right guesses earlier will be among the survivors and will all carry the same 'ancestral genes'. We therefore just keep a record of the family tree (the bit-guess sequence) of each survivor, and can trace back to find the transmitted bit-stream, although we have to wait at least 5 generations (bit periods) before all survivors have the same great great grandmother (who guessed right five bits ago). The whole point is that because the scoring system is based on the running total, the decoder always gives the most accurate guess, even if the received pattern is corrupted, although we might need to wait a bit longer than 5 bits for the best answer to become clear. In other words, the Viterbi decoder corrects errors.

The longer we wait, the more accurate it is. I chose a decoder delay of 4 times the time spread, or 20 bits. We now have a 25 bit delay from one end to the other, (800mS), giving a round-trip delay to a two-way contact of 1.6 seconds. I think this is about the limit before it becomes a nuisance. In any case, the decoder could change to trade performance for delay without incompatibility.

QPSK on the Air

PSK31 operators find QPSK can be very good but is sometimes disappointing. In bench tests with white noise, it is actually worse than BPSK, confirming the simulation work mentioned earlier, but in conditions of burst noise, improvements of up to 5 times in the character error-rate have been recorded. This performance doesn't come free, however. Apart from the transmission delay, which can be a bit off-putting, QPSK is twice as critical in tuning as BPSK. A QPSK signal will start to decode wrong when the phase shift is 45° out, and that will be the case when

Screenshot of the control panel of the PSK31 Windows program, with the tuning display showing a slightly noisy QPSK signal, and showing the fine-tuning controls of the receive and transmit audio tones.



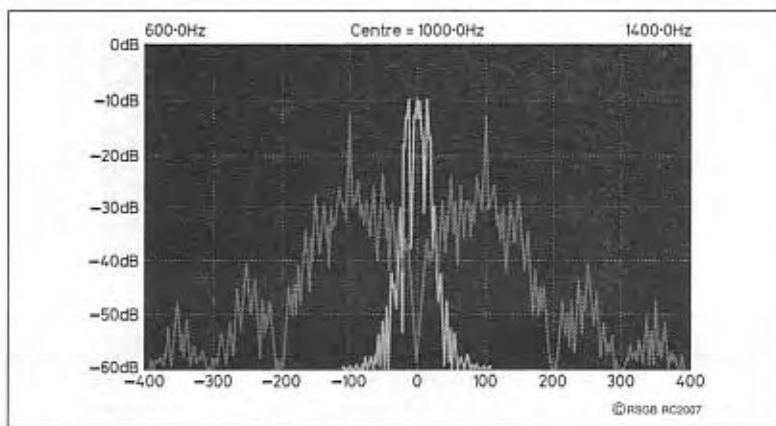


Fig 5.24: Comparison of PSK31 spectrum with 100 baud, 200Hz shift FSK.

the tuning error is only 3.9Hz. This could be a problem with some older radios. What tends to happen is that contacts start on BPSK and change to QPSK if it is worth doing and if there is no drift. There is one aspect of QPSK that has to be kept in mind - it is important for both stations to be using the correct sideband - on BPSK it doesn't matter.

Extending the Alphabet

In the UK, our computer keyboards have a Pound sign above the figure 3, and many people will have noticed that they can't reliably send Pound signs over the Internet, for example. This is because the Internet uses the 128-character ASCII alphabet, and the Pound sign is not part of that set but part of the ANSI character set which has an additional 128 characters and symbols. PSK31 as described so far is the same as Internet. It's a small problem in the UK, but much more of a nuisance in other parts of the world where characters like the German umlauts, French accents, and Spanish tildes are also missing from the ASCII character set. Luckily, with PSK31, the Varicode alphabet is very easy to extend without creating incompatibility with earlier versions. Since the Windows® operating system uses ANSI, and most PC programs are now written for Windows, I have recently extended the PSK31 alphabet in a Windows version.

In the basic PSK31, if there was no '00' pattern received 10-bits after the last '00', the decoder would simply ignore it as a corruption. If now I let the transmitter legally send codes longer than 10 bits, the standard decoder will just ignore them and the extended decoder can interpret them as extra characters. To get another 128 varicodes means adding more ten-bit codes, all the eleven-bit ones, and some twelve-bit codes. There seemed little reason to be clever with shorter common characters, so I chose to allocate them in numerical order, with code number 128 being 1110111101 and code number 255 being 101101011011. The vast majority of these will never be used. It would not be a good idea to transmit binary files this way!

TABLE 2

The convolutional code.

00000	2	01000	0	10000	1	11000	3
00001	1	01001	3	10001	2	11001	0
00010	3	01010	1	10010	0	11010	2
00011	0	01011	2	10011	3	11011	1
00100	3	01100	1	10100	0	11100	2
00101	0	01101	2	10101	3	11101	1
00110	2	01110	0	10110	1	11110	3
00111	1	01111	3	10111	2	11111	0

Summarising

I have tried to identify some of the characteristics of modern HF data-transmission modes that have contributed to the decline in 'live QSO' operation, unlike traditional RTTY which is still widely used. By concentrating on the special nature of live-QSO operation, a new RTTY mode has been devised which uses modern DSP techniques and takes advantage of the frequency stability of today's HF radios. The bandwidth is much narrower than any other telegraphy mode. Fig 5.23 shows the spectrum occupied by PSK31 and Fig 5.24 compares this to the bandwidth of standard FSK.

At the time of writing (Nov 1998) PSK31 is available for the Texas TMS320C50DSK written by G0TJZ, the Analog Devices ADSP21061 'SHARC' kit written by DL6IAK, and the Motorola DSP56002EVM written by myself. One or two people are making promising progress with sound cards, and there is scope for implementation of PSK31 with traditional hardware modulators and demodulators. The two tables contain sufficient information to define PSK31 for those that want to try it themselves. All the ready-to-run software and news of the latest developments and activity can be found on the PSK31 Internet web page at <http://bipt106.bi.edu.es/psk31.html>

Referring to Table 2, the left column contains the 32 combinations of a run of five Varicode bits, transmitted left bit first. The right column is the corresponding phase shift to be applied to the carrier, with 0 meaning no shift, 1 meaning advance by 90, 2 meaning polarity reversal and 3 meaning retard by 90. A signal that is advancing in phase continuously is higher in radio frequency than the carrier.

As an example, the 'space' symbol, a single 1 preceded and followed by zeros, would be represented by successive run-of-five groups 00000, 00001, 00010, 00100, 01000, 10000, 00000, which results in the transmitter sending the QPSK pattern .. 2,1,3,3,0,1,2, . .

Note that a continuous sequence of zeros (the Varicode idle sequence) gives continuous reversals, the same as BPSK.

Latest Information and Software

For more on PSK31, including software to run on sound-cards, see: aintel.bi.chu.es/psk31.html.

6

Test Equipment

Simple Test Equipment for 136kHz

By Peter Dodd G3LDO

Much of the work carries out on 136kHz is experimental. It is important that measurements are made and recorded as you set up or improve your 136kHz station.

The most important item of measuring equipment is the moving coil meter, which with appropriate circuits can be set up to measure DC voltage and current and RF voltage and current.

RF Current Meter

The most common method of measuring the output of the transmitter is to use an RF current meter. The thermo-couple RF meter is also quite useful but has the disadvantage of being non-linear and scale rather cramped on the low current section of the scale – at just the point where it would be useful if it were expanded. Also if the RF current to the thermo-couple RF meter exceeds the FSD of the meter the RF thermo-couple unit in the meter will burn out.

A much more useful RF meter was described by Dennis Walker G3OLM. Although the description that follows is for a current meter with an FSD of 1A the information given by G3OLM will enable you to construct a meter with any current range. An instrument with a switched selection of ranges would be very useful.

A system for measuring current should present a low resistance so as to ensure minimum disturbance to the circuit conditions. Low resistance is, of course, a relative term and in the context of measuring current on aerial wires and transmission lines it is the effective characteristic impedance which determines what is an acceptable value. This characteristic impedance will normally lie in the range 50 to 1000 ohms. As long as the current measuring system has an input resistance of not more than 0.5 ohms there should be little problem in disturbing the conditions by introducing the system.

With the wide availability of small ferrite rings it is possible to take a small current sample in a precise proportion. Rectifying the sample and applying the output to a moving coil meter provides a predictable and reliable method of current indication which will tolerate large overloads, is linear scaled and will respond quickly. Let's design a system to provide a full-scale indication of 1A when using a 100 μ A moving coil meter.

The essential property of a current transformer is that the ampere-turns on the primary are precisely balanced by the ampere-turns on the secondary. This is arranged so

that with a single turn on the primary side and 50 turns on the secondary we can expect precisely 20mA to be available to circulate in the secondary for each amp of primary current. In dealing with toroid ring cores a 'turn' simply means a pass through the central hole – it does not need to be complete.

The other necessary condition is that the secondary load must be low enough to allow the current to circulate, otherwise we do not have a current transformer! A value of 470 ohms is suitable as this will reflect into the primary an equivalent resistance of 470/2500 or 0.18 ohms which is low enough to meet the conditions outlined above (resistance is transformed according to the square of the turns ratio).

With 20mA circulating in 470 ohms we will have 9.4 volts RMS available for rectification, corresponding to 13.2 volts peak. We can expect to lose approximately 0.5 volts at the detector diode leaving a DC voltage of 12.7 to drive the moving coil meter. To obtain full-scale deflection of 100 μ A with 12.7 volts requires a total resistance of 127k Ω . The meter itself will contribute about 800 ohms so a practical 120k resistor will fit the bill with negligible error. Finally the effective resistance of the detector circuit will be much greater than the 470 ohms so there will be negligible error due to loading here. A circuit of the current meter is shown in Fig 6.1.

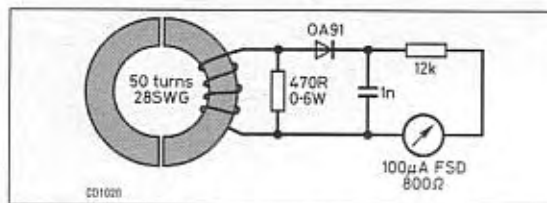


Fig 6.1: Circuit of an RF clamp-on current meter with a FSD of one amp.

Tuned RF Detector for LF

This instrument is very useful, particularly in the early stages of developing an LF station when the characteristics of a transmitting loading coil and antenna combination are unknown.

You can use a signal generator and this tuned detector to determine the resonance of your transmitting loading coil and antenna. Couple the signal generator into the loading coil, either by tapping on to the coil or by using several loop turns coupling. Couple the wire antenna of

the tuned detector to the main antenna by twisting the wires together. Turn the signal generator to maximum output and rotate the frequency dial of the signal generator and the tuned detector until some indication occurs on the meter.

Track the signal generator dial and the tuned detector dial across the frequency range – resonance will be indicated by a sharp increase in tuned detector meter reading.

Many signal generators have low output even at the maximum output setting. Check this by connecting the output directly from the signal generator to the tuned detector. If the tuned detector does not indicate when it is tuned to the signal generator a small amplifier may be necessary. Something like the two BC549 stages of DJ1ZB's exciter (without the filters) should be suitable.

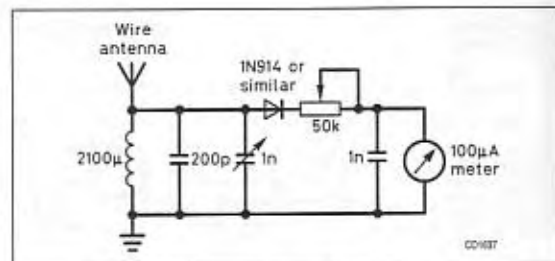


Fig 6.2: A tuned detector for LF. The 2100µH coil can be constructed on a 20mm x 100mm former with the 690turns of 22SWG wire wound over a length of 80mm. The coil size is not critical provided it is resonant on the 136kHz band with some space on the tuning dial above and below. The 50k resistor can be a variable carbon or Cermet pot to give some variable attenuation.

The ScopeMatch Tuning Aid

By Jim Moritz M0MBU (with illustrations by Lech G3KAU)

This device is used to simplify the tuning of an LF antenna fed with 50Ω coax. It is basically an SWR bridge without the detectors and meter as shown in Fig 6.3. With this device the current and voltage amplitude and phase relationships can be monitored on a dual trace oscilloscope to establish a matched condition.

Fig 6.3: Construction and circuit of the ScopeMatch. Note that the coaxial cable screen within the box is grounded on one side only.

Construction

The circuit and construction is shown in Fig 6.3. Any high permeability 18mm diameter ferrite toroid ferrite core can be used for the transformers although a 3C85 core is

ideal. The one in the prototype came from an old SMPS common mode choke. Do not use iron dust cores.

T1 secondary comprises 50 turns of enamelled wire. The primary is a single wire passing through the middle of the toroid, as used in an SWR bridge. Power handling is not really an issue.

T2 uses the same the same construction, but in this case the 50 turns are on the primary, with a single loop of wire to the output connection. Note that the 50-turn winding has the full transmitter output voltage across it, so the winding has to be well insulated from the core to withstand a few hundred volts of RF. The small core worked fine at the 400W level, but saturated with 600W. If high power is contemplated it would be advisable to use a larger core for T2.

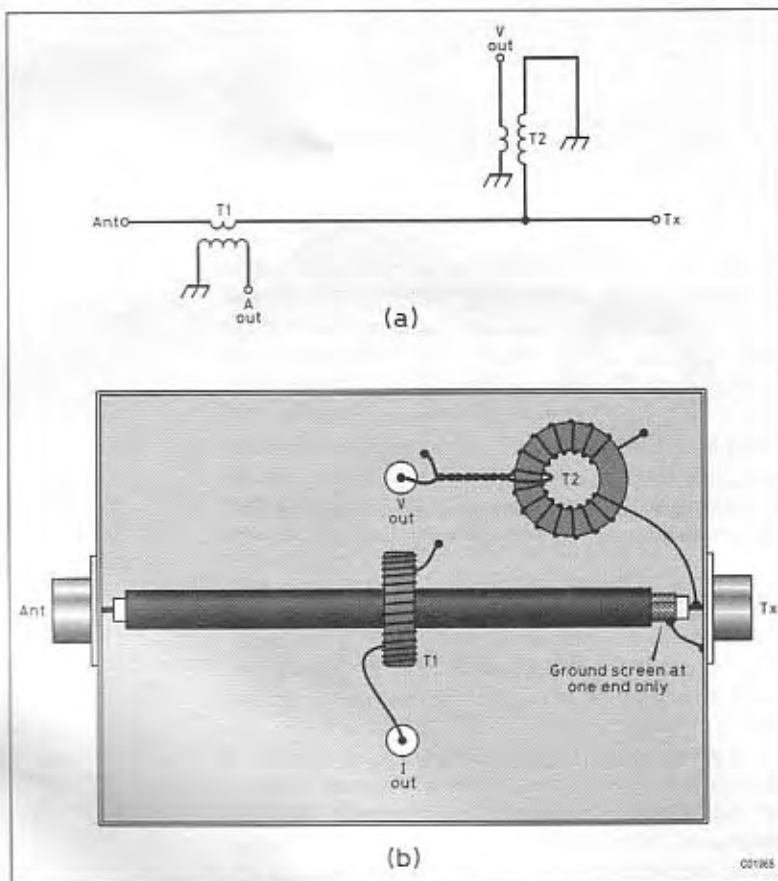
Operation

T1 is a 1:50 current transformer, which samples the current at the Tx output and together with the 50Ω resistor the scale factor is 1V = 1A. T2 is a 50:1 voltage transformer which samples the output voltage, 1V out = 50V at Tx output.

The oscilloscope is set for the same volts/division on both channels; see Fig 6.4. The scale factors are chosen so that when the antenna system is resonant at 50 ohms (or a 50Ω dummy load is used), both voltage and current traces are identical. If the load is inductive, the current waveform will lag the voltage; if capacitive it is the other way round. Getting the antenna resonant is just a matter of adjusting the loading coil until the two waveforms are in phase. Once the antenna is resonant, if the current waveform is bigger than the voltage waveform, the load is less than 50 ohms, and if smaller the load is greater than 50 ohms. You can calculate the actual R by measuring V and I off the screen and using Ohm's law.

I have found this gadget very useful both for setting up an antenna, and while operating it takes out most of the guesswork that occurs when using SWR bridge circuits. You can also tune up on low power – 1W is required at my location owing to the signal from nearby MF broadcast stations – but elsewhere 100mW or less is OK.

T2 can be replaced with a capacitive divider. This com-



prises a 100p capacitor connected to the inner of the coaxial cable at the 'Transmitter' socket in series with a 5000p capacitor connected to ground. The centre of the capacitive divider is connected to the voltage-sampling socket. The 100p capacitor must be rated to take the transmitter output voltage.

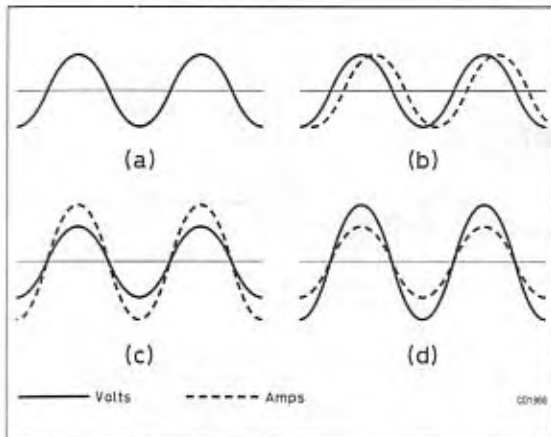


Fig 6.4: (a) Perfect match. Current and voltage waveforms have the same phase and amplitude and only one trace discernable. (b) Antenna off tune and inductively reactive (voltage leads current); (c) Antenna resonant but resistive component of impedance low (voltage 25V, current 1amp, 25ohms) (d) Antenna resonant but resistive component of impedance high (voltage 50V, current 0.5amp, 100ohms).

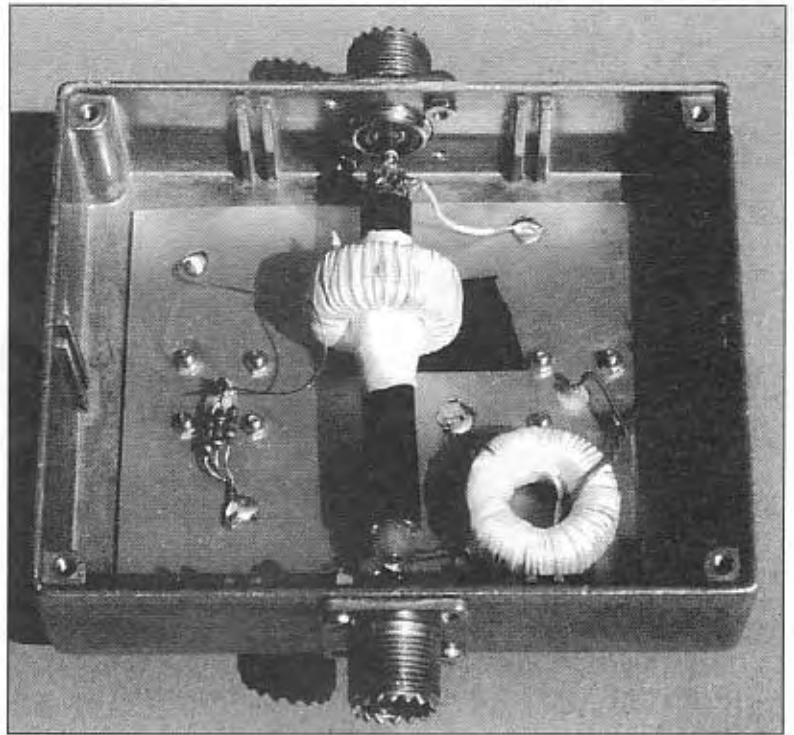


Fig 6.5: The ScopeMatch device as constructed by G3KAU, shown from the underside. The voltage and current sampling connections to the oscilloscope are made via two BNC connectors.

Field Strength Meter for the 137kHz Band

By Dick Rollema PA0SE

The power radiated by an antenna is equal to the radiation resistance multiplied by the antenna current squared. Measurement of antenna current is described earlier. The unknown factor is the radiation resistance. Computer programmes for antenna simulation can produce a value for the radiation resistance but proper modelling of the antenna is not always easy. Another problem is the influence of the earth. The ground constants are seldom known and even if they are it is not certain that the computer program applies them in the correct way.

A more reliable way of determining radiated power in the 137kHz band is by measuring the field strength near the station but outside the near field region. A distance of 1km is probably sufficient to reduce the influence of the near field on the measurement sufficiently and 2km is definitely safe.

At such a distance we are in the far field of the antenna but near enough so that the field strength does not depend on the type of ground. When a strength of the electric field of E mV/M is measured the radiated power follows from a simple equation:

$$P = 0.0111(E \times d)^2 \quad (1)$$

Where P = watts, E = mV/m and d = kilometres
(x = multiplication)

The equation produces the power really radiated by the antenna, in other words the power 'dissipated' in the radiation resistance.

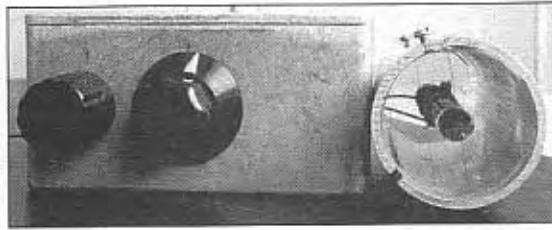
Note that this is not the same as ERP. By definition ERP is the fictitious power to be fed to a *half wave dipole*

in free space that produces the measured field strength. As ON7YD has pointed out, a short vertical (and our 137kHz antennas are always short) has a theoretical gain of a factor 1.83 (2.62dB) over a half wave dipole in free space. So if you want to know your ERP multiply the power given by equation (1) by 1.83 (or add 2.62dB). But apart from a regulations point of view I see no advantage in using ERP.



Fig 6.6: The PA0SE LF Field Strength Meter.

Fig 6.7: Side view of the instrument showing the construction of the antenna. The ferrite rod is held in place in the tube by two discs of Perspex.



The actual power radiated by the antenna is what counts. Most field strength meters do not measure the electric but the magnetic component of the electromagnetic field. But this is no problem because in the far field of the antenna (where we measure) there is a fixed relation between the electric and the magnetic field components:

$$E/H = 120 \times \pi \text{ ohms} = 377 \text{ ohms} \quad (2)$$

Where $E = V/m$ and $H = A/m$.

The portable field strength meter to be described is a direct conversion receiver with two audio output signals. One is fed to headphones for tuning the meter to the signal to be measured. The other output feeds a digital multimeter. The voltage indicated by the DVM has a linear relation to the field strength as shown in Fig 6.6. The meter is calibrated so that a field strength of 5mV/m produces a reading of 1V on the DVM. The instrument can be tuned over the range 135.530 to 139.296kHz. This includes DCF39 on 138.82kHz which is useful for comparison purposes.

Description of the Instrument

The antenna is a ferrite rod from a broadcast receiver with the original long wave coil in place. The rod is centred in an aluminium tube of 32mm internal diameter and 145mm length as shown in Fig 6.7. The aluminium tube has a slot cut along its length to prevent it becoming a short-circuited turn.

The electronic circuitry is housed in a diecast aluminium box of 120 x 95 x 61mm and the circuit is shown in Fig 6.8. The antenna is tuned by capacitors C1, C2 and C3. C2 and C3 were selected so that C1 can tune the antenna to the centre of the band. R4 was added to widen the frequency response so that it is sufficiently flat over the range of interest.

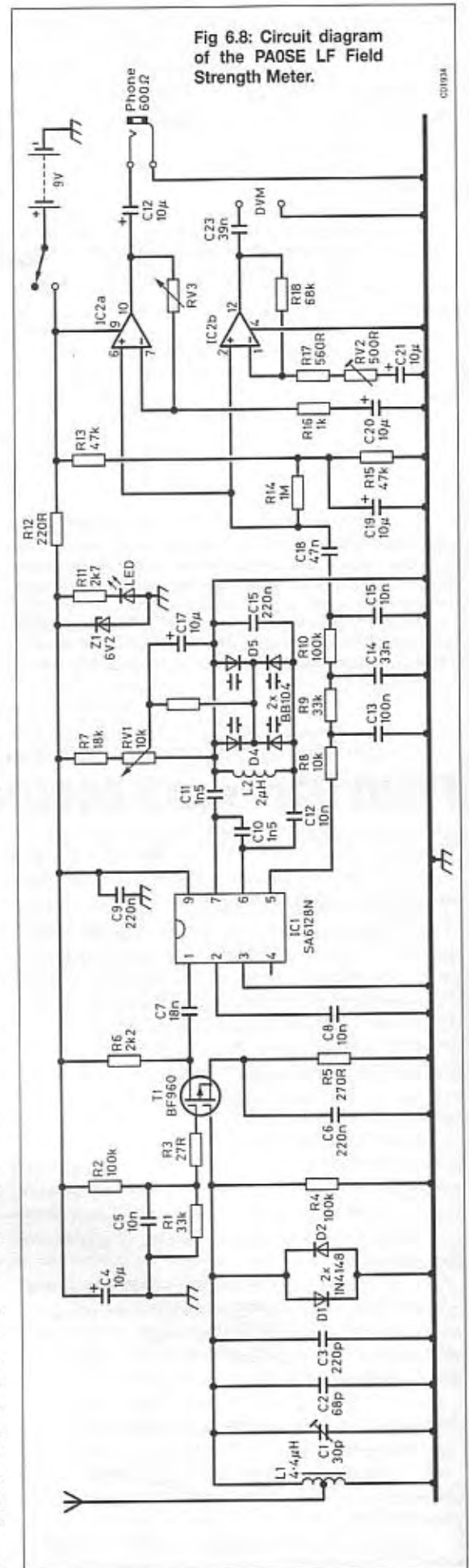
D1 and D2 protect the instrument when it is used close to a high power transmitter. T1 amplifies the signal without loading the antenna circuit. It is important in a direct conversion receiver that the signal from the local oscillator cannot reach the antenna circuit. The RF amplifier, including C7, is therefore mounted on a separate piece of copper clad circuit board (not etched). Where insulated tie points are required small islands of PCB material is fixed to the board with instant glue.

The mixer and output stages are built on a second board. These boards are fitted into the diecast box with the battery and controls as shown in Fig 6.9.

The mixer is an IC type SA612BN, which incorporates an oscillator circuit. The values of C10, C11 and C16 were dictated by the choice of coil L2. I used a junkbox 2mH HF choke. C10 and C11 are the usual capacitors found in a Colpitts oscillator. C12 isolates L2 from the DC on pin 6 of the mixer. The BB104 type dual varicaps also came from the junk box and two of these in parallel were necessary to obtain the required tuning range. The lower frequency was set by selecting C16. The upper limit was found to be a bit high and this was corrected by adding R7. RV1 is the tuning control.

Selectivity of a direct conversion receiver is determined

Fig 6.8: Circuit diagram of the PAOSE LF Field Strength Meter.



by a low pass filter in the audio path. A high degree of selectivity is required here because of the extremely strong station DCF39 at 138.82kHz, only 1020Hz above the upper limit of the band. I use a RC-filter with three sections, each having a time constant $RC = 1\text{ms}$. At first resistors and capacitors of the same value were used in the three sections. Later I realised that a better response is obtained when the loading of a section on its preceding one is decreased and this resulted in the values seen in the circuit diagram. The lower limit of the frequency response is set by the time constant $RC = (R17 + RV2) \times C21$ respectively $RC = R16 \times C20$. The response of the metering circuit shows a maximum at about 36Hz and is 3dB down at 16 and 88Hz.

The output of the low pass filter is fed to the two sections of a dual op-amp type UA747. The upper op-amp feeds the headphones. Volume is controlled by RV3 in the feedback path.

The lower op-amp feeds a digital multimeter that must be capable of measuring AC in the millivolt range up to about 2V. Preset resistor RV2 is adjusted when calibrating the instrument. I choose to make the audio output, as indicated by the DVM, 1000 mV when the instrument is placed in a field of 5mV/m. The reading is linear up to about 10mV/m maximum (2V on the meter).

The instrument is energised using a 9V battery. The LED is a small one that gives a clear indication of the instrument being switched on when drawing a current of only 2 mA.

To make the gain of the RF amplifier and mixer independent of battery voltage the supply for these stages is stabilised at 6.2V by a zener diode. R12 was selected so that the zener keeps control for battery voltages down to 7 V. There is no need to stabilise the supply for the op-amps because their gain is controlled by negative feedback and therefore hardly depending on the supply voltage. The instrument draws about 17mA from a new battery.

Calibration

To calibrate the field strength meter the instrument must be placed in a magnetic field of known strength. This can be produced by a pair of so called 'Helmholtz coils'.

In the 19th century German scientist Helmholtz found, by computation, that a homogeneous magnetic field can be produced by placing two circular one-turn coils of radius r metres parallel to each other at a distance of r metres and with their axes coaxial. When a current I is made to flow through each of the coils in the same direction a high intensity homogeneous field of

$$H = I / (1.40 \times r) \quad (3)$$

Where $H = \text{A/m}$; $I = \text{amps}$; $r = \text{metres}$

is generated between the coils. I constructed a pair of Helmholtz coils with $r = 0.292\text{ m}$ as shown in Fig 6.10.

The coils are connected in parallel and in series with a 50Ω resistor. At 137kHz the reactance of the coils is so low that it can be neglected against the 50Ω resistor. Therefore when the coil pair in series with 50Ω is connected to a signal generator with U volt output the current I through the coils in parallel is: $I = U / 50$. (I in amps). Note that each of the coils carries half of that current.

The calibration setup was checked using a single turn 10cm diameter coil, connected to a selective level meter via a length of coaxial cable. I had calculated that with 1 volt applied to the Helmholtz coils in series with 50Ω a voltage of 208 microvolts should be induced in the 10cm

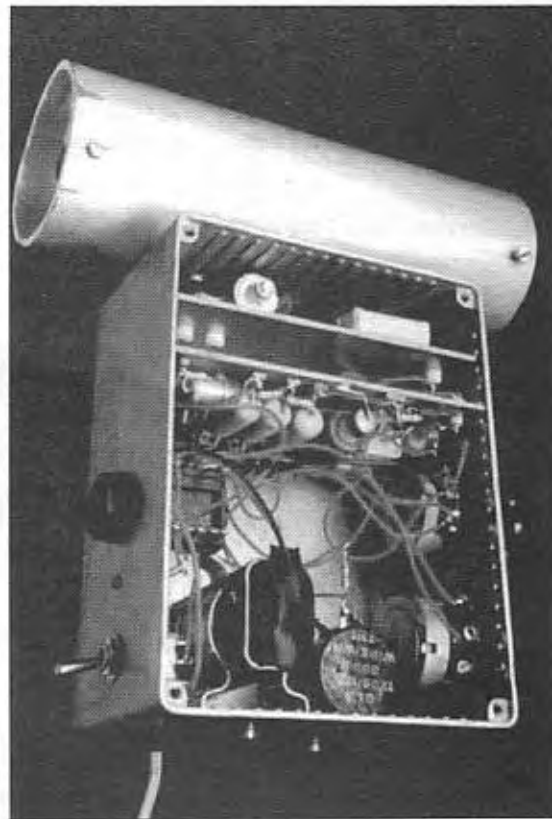


Fig 6.9: General construction of the LF field strength meter.

coil when held between the Helmholtz coil pair. I measured 210 microvolts; almost too good to be true. The calibration setup was double checked and found to be correct.

The field strength meter was required to produce a reading of 1000mV in a field strength of 5mV.

Equation (2) was applied and it was found that the corresponding magnetic field component was $H = 13.3 \times 10^{-6} \text{ A/m}$.



Fig 6.10: Field strength meter calibration arrangement using Helmholtz coils.

Using equation (3) the generator output voltage was found that would result in the wanted field between the Helmholtz coils. The field strength meter was placed between the coils and RV2 adjusted for a voltage of 1000mV on the digital multimeter. That completed the calibration.

Measuring field strength

Try to find an open space at least one kilometre from the transmitter. Keep the antenna of the meter horizontal and tune the signal to zero beat. Now slowly increase or decrease the tuning slightly for a maximum reading on

the DVM. Whether or not you can hear the beat note of about 36Hz depends on the quality of your headphones. Turn around slowly to find the position for maximum signal. Now walk around a bit. If the reading varies the field is probably being distorted a metallic fence, a lamppost or underground cables or pipelines (on 137kHz the waves penetrate tens of metres into the earth) When a constant indication is found, multiply the reading in volts by five to obtain the field strength in mV/m.

Find the distance to the transmitter on a map or by other means, such as GPS. Apply equation (1) to find radiated power. Multiply by 1.83 if ERP is required.

Measurement of Antenna System Impedance at LF

By Peter Dodd, G3LDO

The feed impedance of an LF antenna fed against ground can vary considerably, depending on antenna size, ground characteristics and loading coil construction. Although a suitable matching transformer can be arrived at empirically the design can be simplified if the antenna feed impedance is known.

An impedance value will also give you some idea of how good your ground system is, once you have estimated the coil loss.

The 3-meter Impedance Measuring Method (3m Z Bridge)

This method of measuring impedance is very simple. I have used this method of measuring impedance for the last 20 years on HF and it is described in more detail in *The Antenna Experimenter's Guide*. What follows is the simple modification to original design and a short description of how to use it on an LF antenna.

The 3 meter impedance measurement technique com-

pares the unknown impedance with a fixed standard impedance, and the ratio is indicated by three voltmeter readings, see Fig 6.11. In this variation of the method, five readings are made. One of the additional readings allows in-place calibration of the reference capacitor and the second permits several solutions for the unknown impedance, thus giving an indication of the random errors that may be present in the data.

The fixed standard impedance comprises a resistor and capacitor. An RF excitation voltage, at the measurement frequency, is applied to Z via R and C. The voltages across R and C are measured together with the input voltage E_A , the voltage across Z, and the voltage across Z plus C. The excitation level is adjusted to a specific level, eg $E_r = 5$ volts, then all the other voltages are measured.

Impedance values are derived from the voltage data by calculation, however, this may be simplified in one of two ways:

- By using a graphic method shown in Fig 6.12. The voltage readings obtained from the instrument are plotted on graph paper using a pair of compasses and a ruler. The complex impedance is identified as the point where a set of three arcs intersect.
- By using a computer.

The RF voltages are measured using diode probes, selected by a switch. These probes measure peak volts and require a high impedance voltmeter; a digital voltmeter is ideal. The full circuit is shown in Fig 6.13.

For greatest accuracy the value of R and the value of C need to be appropriate to the range of impedance and frequency respectively, of the measurements being made. The impedance range of most amateur LF antennas is around 40 to 90 ohms. The original value of R used in the design of the 3-M box R was 50 ohms because the method was used to measure impedances in 50Ω systems, however this still turns out to be suitable for our purposes.

The reactance of C can also be 50 ohms. In practice a reactance value for C of between 25 ohms and 100 ohms, will give reasonable results

A low power transmitter with a variable power output was originally used as excitation source. An attenuator was used at the input so the transmitter was isolated from the variations of unknown impedance as the transmitter frequency is varied during a series of measurements. This attenuator is dispensed with for the LF version.

Fig 6.11: Block diagram of 3 meter impedance bridge showing the voltage measurement points.

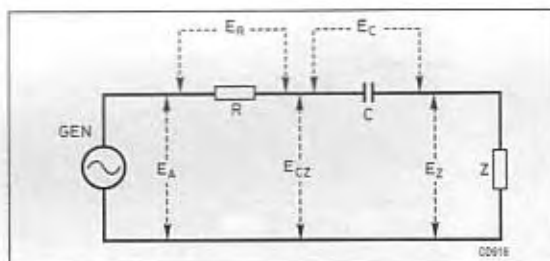
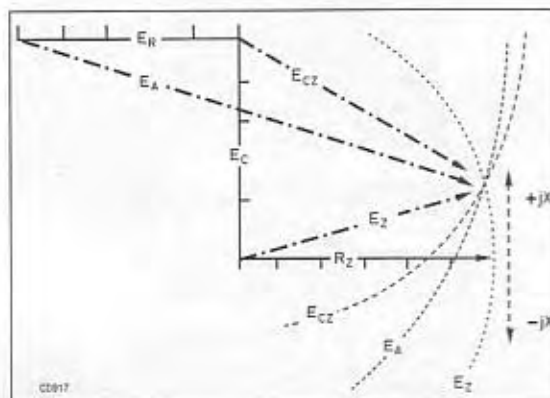


Fig 6.12: Method of determining impedance from measured voltages; see Fig 6.1.



For the LF measurements I used a signal generator with a small audio amplifier as the excitation source. It is very important that the harmonic output from the excitation source is kept as low as possible.

A general view of the impedance matching set up is shown in Fig 6.14.

Using the 3M Z Bridge

Connect the unknown impedance, excitation source and digital voltmeter to the 3-Meter test bridge as shown in Fig 6.15.

If you are using the standard Marconi antenna measure the feedpoint impedance from the bottom of the loading coil and earth. There is no DC path to ground when this instrument is connected so do not use if there is a lot of rain static around. If you are measuring the impedance of a large antenna it might be wise to use a 1:1 transformer, see Chapter 4.

This method is used to extract the impedance values from the voltage data:

- Set the switch to read E_r .
- Increase the output from the signal generator until the voltmeter reads 5 volts; record this on a notepad as 50.
- Note the voltages at the other switch positions, (12V being recorded as 120 and 6.3V as 63 etc); then check the stability of the transmitter power output by checking position E_r again. The readings should be repeated if the excitation level has drifted more than about 0.1V.

Graphic Method of Extracting Impedance from Data

All that is required is a sheet of linear graph paper, a pencil and a pair of compasses. The method is illustrated in Fig 6.16.

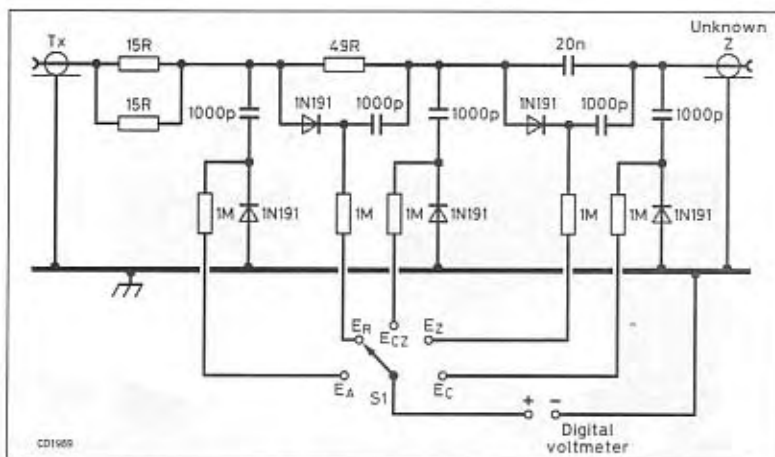
- 1 Draw a horizontal line on the graph paper whose length is equal to E_r .
- 2 Draw a vertical line down from the right-hand side of line E_r whose length is equal to E_c .
- 3 Place the point of the compasses on [1] and draw an arc whose radius is equal to E_a . Repeat for [2] and [3] for radius values E_{c2} and E_z respectively.
- 4 Mark the point at which the arcs intersect. An exact intersection of all three arcs is not always possible due to the errors in the data.

A horizontal line from the reference point to the arc intersection gives the resistive value of impedance. A vertical displacement up or down from the resistive line gives the value of the inductive or capacitive reactance respectively. If the arcs do not intersect take the centre of the 'triangle' formed by the non-intersecting arcs as the impedance point.

Extracting Impedance from 3-M Data using a Computer

A program is available from the RSGB web site for extracting impedance from 3-M data using a computer. When the program is run you will be prompted for items of data. The program prints the solution as soon as the last item of data is entered. In addition the program will report errors, which will give an indication of data reliability.

The program will amend any data resulting in a small geometric non-intersection. It will also inform you of any change in any item of data required to effect this intersection. The program will make 10 attempts to correct a non-



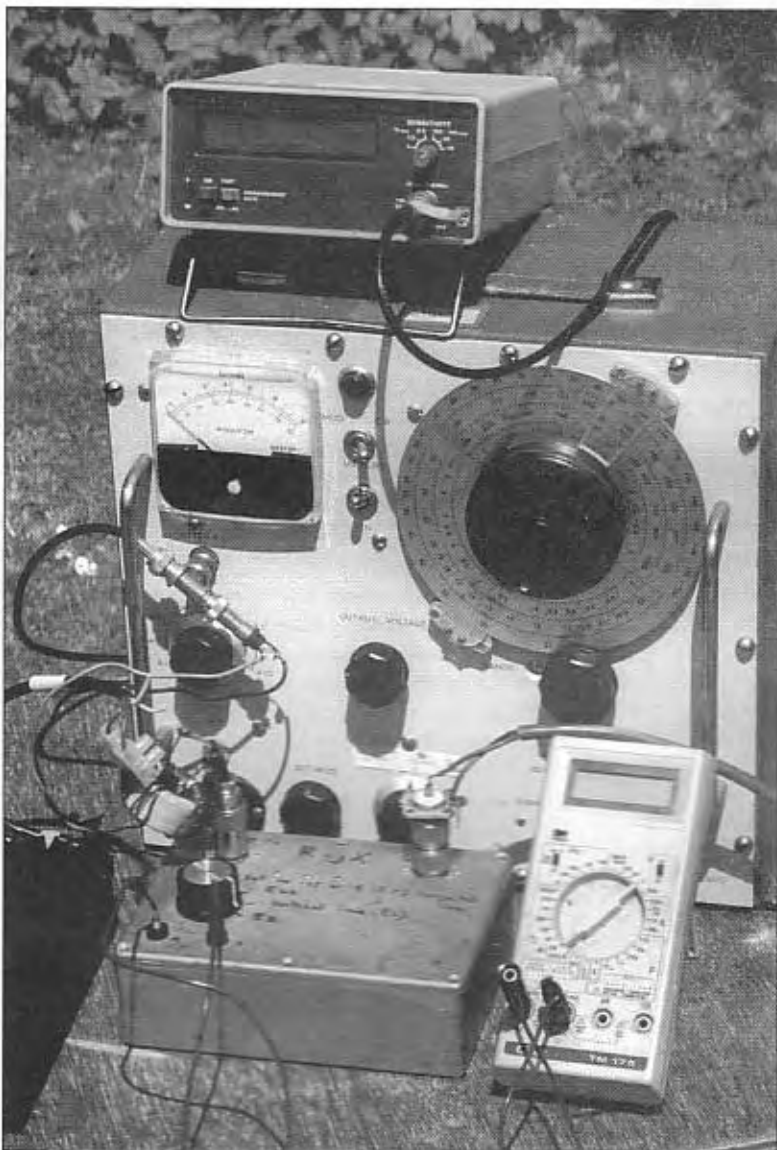
intersecting error. If the data cannot be corrected during these attempts the program will report this and not give a solution.

Fig 6.13. Three-M circuit diagram.

Excitation level

You will have noticed that in the procedure for using the 3-M box I have stated that the excitation level should be increased until the voltmeter reads 5 volts across the reference resistor. The reason for this is mainly historical. In the original article W8CGD used the graphic method for

Fig 6.14: Equipment required to make the impedance measurements at LF.



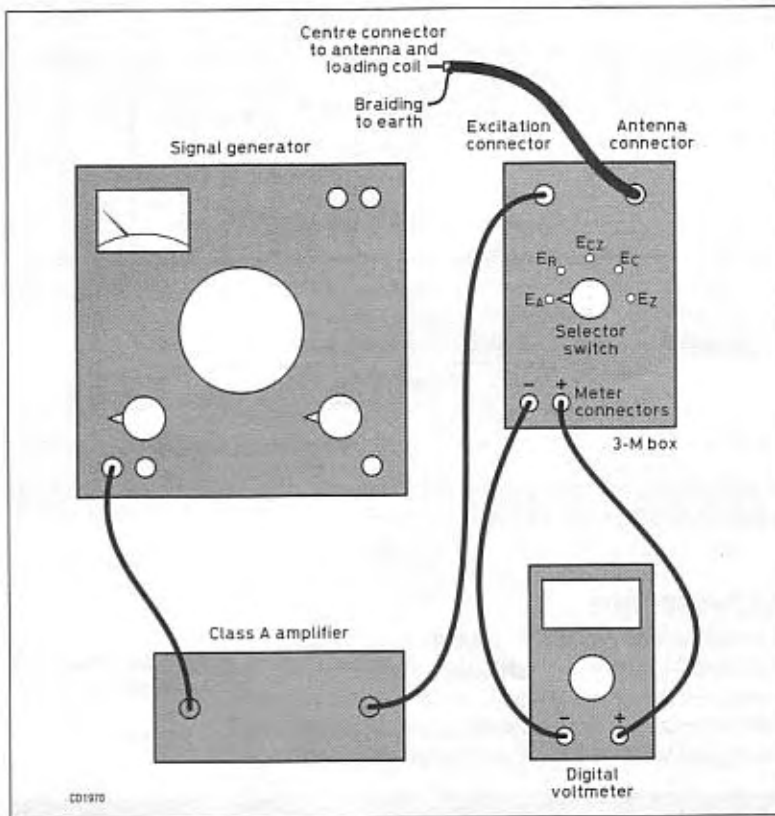


Fig 6.15: The 3-Meter test bridge and connections for antenna impedance measurements.

extracting the impedance from the voltage data, and because the impedance measurements were related to a 50Ω system, 50 ohms became the standard for the fixed reference components and the divisions on the graph paper.

You don't have to use 5 volts, after all it takes a relatively high level of excitation power to achieve this.

Provided all the voltages are in the correct ratio, theoretically the results should be the same.

However, all the computer programs were just software implementations of the graphic method and they will only work with an E_R measurement of 0.5, 5 or 50.

A program is available, which enables the use of other excitation levels. It also normalizes the data for use with programs, which can display a set of data in tabular or graphical form as shown in Fig 6.17.

In practice any excitation level appeared to give good results but errors increased sharply if any one of the measured voltages fell below one volt. This level may depend on the type of diode being used but the program will give a good indication of the errors if the excitation level is too low.

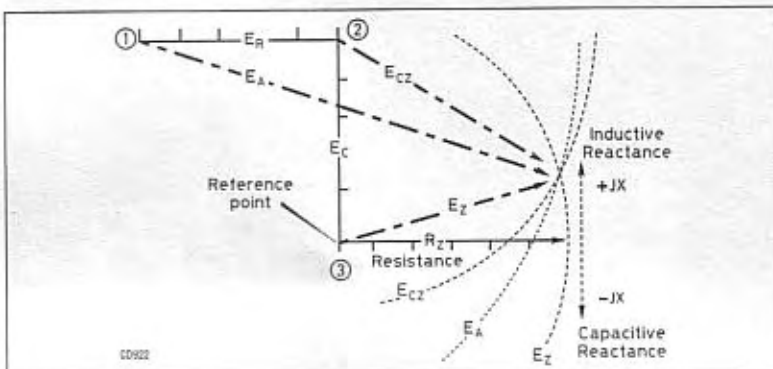


Fig 6.16: Method of extracting impedance from 3-M data using linear graph paper.

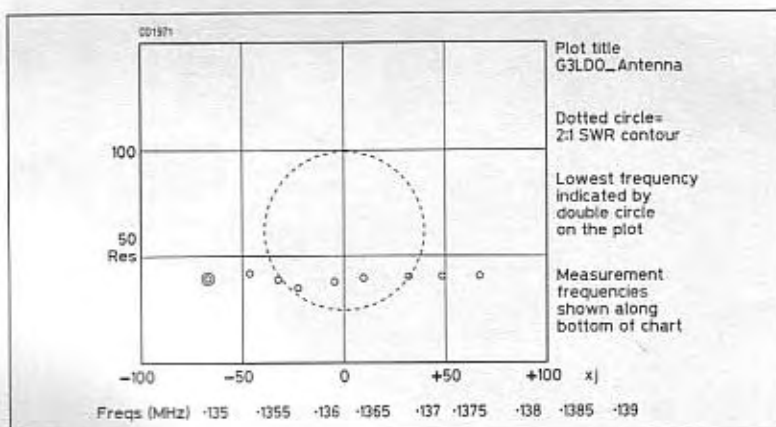


Fig 6.17: Impedance plot of the G3LDO antenna shown in Chapter 4. The variation in reactance is considerable over the measured frequencies. However, the resistance hardly changes because most of the resistance is made up of aperiodic ground losses.



Appendix 1

Most of us, by now, have had some experience of trying to obtain a good RF ground at LF. This was also a problem to early radio engineers trying to obtain highest efficiency from their LF antenna systems. The following article is from *The Year Book of Wireless Telegraphy and Telephony* 1922 and illustrates the amount of effort that was involved in trying to obtain a good earth.

The article is presented in the hope that we may learn from these early experiments. Some of the sentences are

rather long and convoluted, which may be the result of the translation. I have only edited the material where absolutely necessary.

The unit Centimetre is mentioned. This is an absolute unit of measurement and its equivalents (according to the *Admiralty Handbook of Wireless Telegraphy*, 1925) are as follows:

One Farad = 9×10^{11} Centimetres
One Henry = 19^9 Centimetres

The Earthing Resistance of Antennae

By DR ING A MEISSNER

A communication from the Laboratory of the Gesellschaft für drahtlose Telegraphie

For many years one of the most important problems of radiotelegraphy has been the question of the best type of transmitting antenna for high-power work. Prior to 1911 no satisfactory solution had been found. On the one hand L-shape antennae of 1000 to 2,000 metres in length and 50 to 80 metres height had been built, while, on the other, the umbrella-shaped antenna was preferred by many.

The former type was said to have a very pronounced directive effect, and theories have been worked out to prove the superiority of this type of antenna. It had also been proposed to use earth-antenna for transmission purposes.

In order to determine the basis for the design of the long-distance radio stations, which were projected about that time for erection in the German Colonies, the Telefunken Company resolved to carry out extended and more detailed experiments with antennae. These researches were concerned with (1) the radiation from the antenna, and (2) the design of its leading dimensions, viz., the predetermination of its capacity, natural frequency and damping and the influence of the insulation of the metal parts of the masts, Supporting wires, etc. The first experiments carried out at Nauen were followed by tests over longer distances, between Nauen and Jena. The following were the main results obtained at that time, relations which are still generally used in the design of antennae:-

(1) That the statement that the radiation resistance is proportional to h^2/λ^2 is true for all shapes of antennae.

(2) That within wide limits if the mean height is the same, the shape of the antenna has no influence. Umbrella, T and L aeriels are equivalent if the earthing or

counterpoise is the same; but an L aerial is not so good if the length l is greater than four times the height h , on account of the losses in the ground.

(3) The directional effect of an L antenna, when the loading [1] is more than 20 per cent, is practically negligible.

A solution of one problem was not obtained at that time, viz., by what means can the earthing resistance of the antenna be reduced?, and this question became of greater importance during the following years, since the efficiency of the whole antenna, especially for large aeriels, depends mainly on the earthing resistance.

Although the efficiency of the oscillation generator has during the past few years been increased up to 70 to 80 per cent, the antennae have become less effective as longer and longer wavelengths have been used, in order to decrease the absorption and to reduce the interference from other stations. Although the highest practical towers have been used, a radiation resistance of between 0.1 and a little over 0.3 ohm was the maximum obtained at wavelengths of 12 to 20 kilometres, while at the same time the earth resistance has increased to from 1.5 to 5 ohms. Hence the efficiency of the whole antenna is but 5 to 15 per cent. It is obvious, therefore, that the problem of increasing the effectiveness of large antennae can only be solved by a reduction of the earthing resistance.

The following results of research work indicate the best way in which to effect an improvement in the efficiency of large antennae.

In Fig 1 are given curves of the antenna resistance as functions of the wavelength or rather of the factor $\lambda/4h$ for the three chief types of antennae: a normal land aerial, a ship aerial, and the aerial of a long-distance radio station. The curves show that the antenna of a high-power station having a capacity equal to 30,000 centimetres [see note above], has a much smaller resistance (2 to 3 ohms)

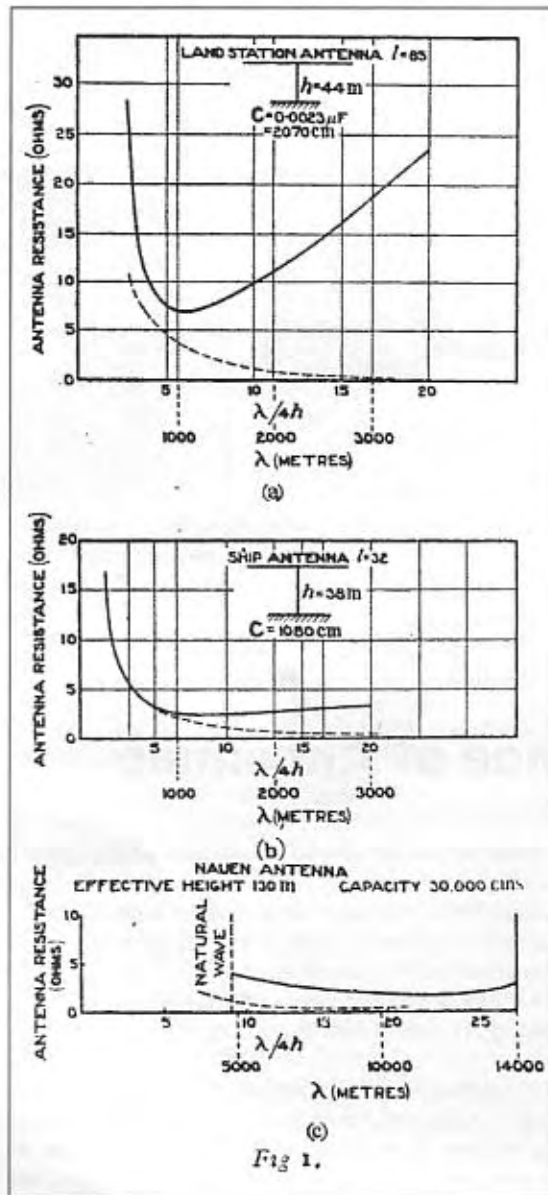


Fig. 1.

than the small T antenna with a capacity of 1,000 centimetres, when both are loaded to the same extent, i.e., the resistance of the antenna decreases when the dimensions are enlarged. (The earth resistances are indicated by the dotted lines.)

In 1913 and 1918, M. Abraham treated this problem

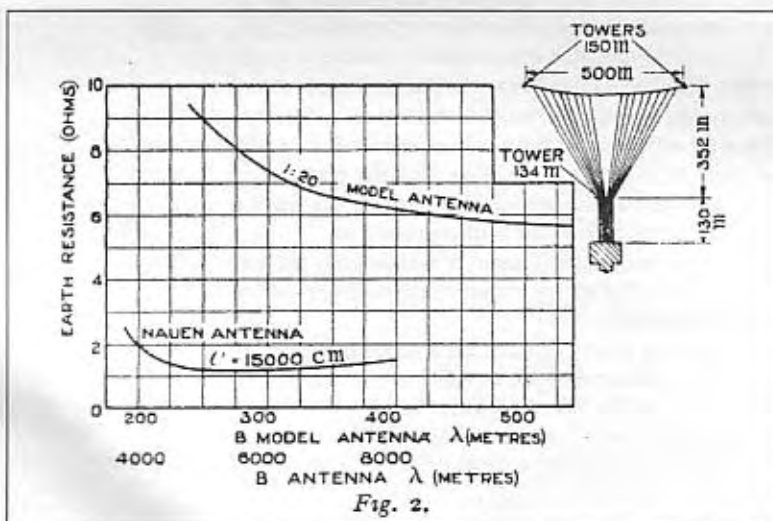


Fig. 2.

theoretically and obtained the following result that the product of the conductivity of the ground and the dimensions of the aerial is a constant. According to Abraham's theory, the earthing resistance of an antenna should be reduced to half if the dimensions are enlarged in the ratio 1:2 and the conductivity of the ground remains the same. The curves in Figs 1a and 1c show that this theory is not rigidly obeyed in practice.

Fig. 2 shows this more plainly. It gives the curves of the true earth resistance of a large fan-shaped L antenna of mean height 80 metres, length 400 metres, breadth at the end 550 metres, and capacity 1,000 centimetres, and also of a model of the same antenna built on the scale of 1:20. It is seen that although the resistance is greatly reduced, it is not in the ratio of one-twentieth, although the ground was the same. It is, however, difficult to be certain whether the ground is identically the same near the earth wires at a depth of 2 to 5 centimetres and at a depth of 40 to 60 centimetres. Abraham's theory can only be applied if all the dimensions of the antenna are changed in the same scale, but it does not apply when only single ones are enlarged while others remain constant.

The question as to whether the earth resistance for a given wavelength can be reduced only by enlarging the antenna is a particularly important one. The curves in Fig. 3 relate to this question and refer to the case of twelve-

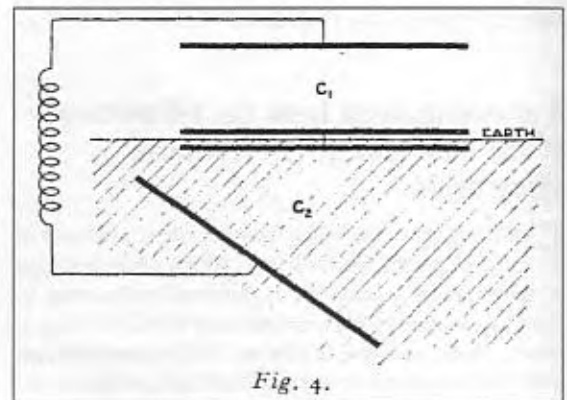


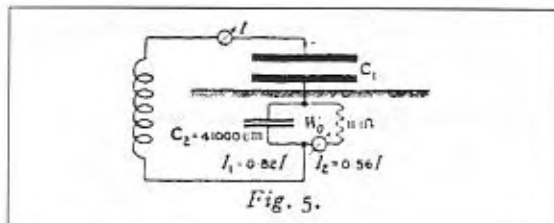
Fig. 4.

wire umbrella type antenna, height 10 metres, length of wires 10, 20 and 35 metres, span 60 metres and with twelve earth wires, each 25 metres long. The curves show the great improvement obtained on long wavelengths when the antenna is enlarged from 10 to 20 metres, the resistance falling more rapidly than proportionally to the increase in size. For example, at a wavelength of 900 metres:

$$\frac{\text{Resistance of 20-metre aerial}}{\text{Resistance of 10-metre aerial}} = \frac{2 \text{ ohms.}}{5.3 \text{ ohms}}$$

The proportionality, however, is not retained when the length of the antenna is increased from 20 to 35 metres, nor when shorter waves are used nearer the natural frequency of the antenna. These values were possibly somewhat influenced by the special conditions of the ground, since part of the area was flooded with water. The rapid rise of resistance indicated by curve i in Fig. 3, as the wavelength is increased is very striking. Howe has assumed that this rise must be a straight line, and consequently due to pure dielectric losses, which are inversely proportional to the frequency, and therefore directly proportional to the wavelength.

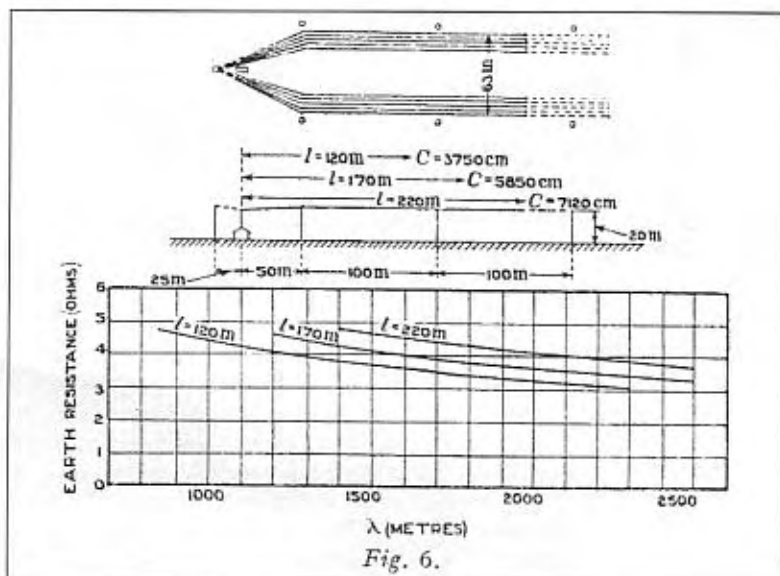
The antenna can be regarded as consisting of two capacities, the loss-free air condenser C_1 (Fig. 4) in series



with the condenser C_2 having a bad dielectric. This arrangement is equivalent to that shown in Fig 5, in which C_2 is represented by a loss-free condenser in parallel with an ohmic resistance. The current I (Fig. 5) increases as C_2 is increased and as the wavelength is reduced. The resultant effective resistance W (i.e., the measured earth resistance) therefore increases as the wavelength is increased and falls with increased capacity C_2 . In general C_2 rises when the antenna capacity C_1 is increased, i.e., W is reduced if the antenna capacity is increased. In this case the condition is that C_2 is increased if C_1 is increased that is, that when the antenna is enlarged, the additional capacities C_1 and C_2 , which are equivalent to the existing capacities are connected in parallel. This is not the case with some shapes of antennae; for instance, with L-shaped antennae the values of W_0 and C_2 can be determined from the resistance curves in Fig 3, and thus the ratio of the currents in both branches can be determined. Fig. 5 gives these current values for a wavelength of 650 metres.

Increasing the antenna capacity is, therefore, one means of reducing earth resistance; but, unfortunately in practice, the fundamental wavelength of the aerial soon becomes comparable with the working wavelength, and this, again, causes the resistance to rise owing to proximity to the natural frequency of aerial. Hence it is difficult to reduce the resistance below 2 to 2.5 ohms, even when the capacity conditions are the most favourable.

This method of reducing the earth resistance is of no use for unsymmetrical antenna, as then the resistance increases when the capacity is increased. Fig. 6 gives resistance curves for L-antenna of different lengths (height of mast 20 metres, breadth of aerial 63 metres, number of wires 10, lengths 120, 170 and 220 metres, capacities 3,750 centimetres, 5,850 centimetres and 7,120 centimetres). From these curves it can be seen that when the capacity of the aerial was increased in the ratio 1:2, the resistance increased by 25 to 30 per cent. (see curves



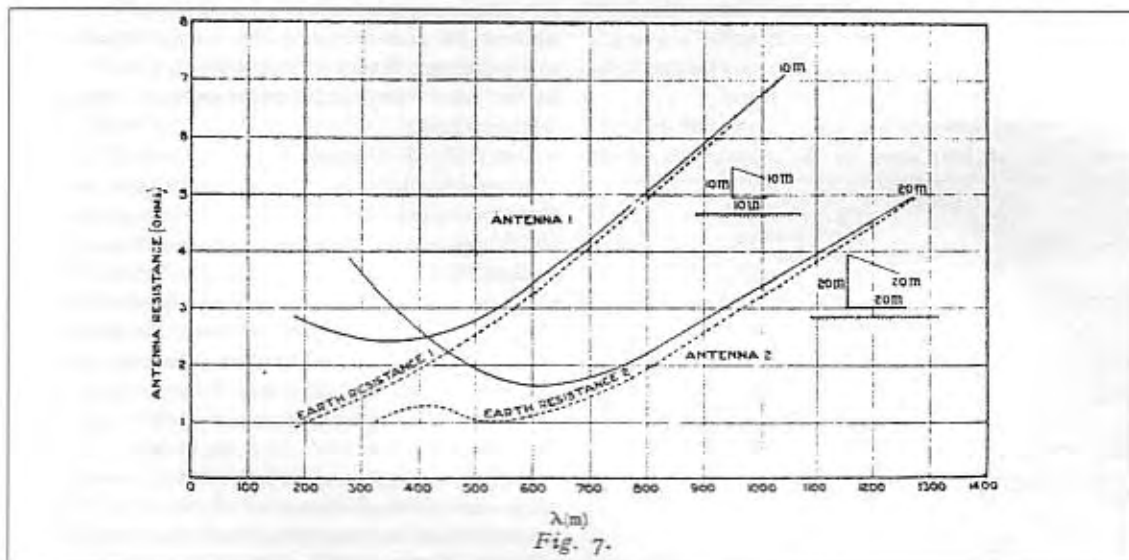
for 120 and 220 metres on wavelength 1500metres). These figures indicate that the additional capacity obtained by increasing the size of the aerial does not add in parallel to the original area; that is, that the capacity of C_0 in the diagram in Fig 5 has been decreased instead of increased, by increasing the antenna capacity C_1 .

The question must now be considered whether a further reduction of the earth resistance is possible by the use of a counterpoise, such as was first introduced by the Telefunken Company about this time.

Fig. 7 shows the change in resistance when all the dimensions of an antenna were altered in the ratio of 1:2 and when a small counterpoise was used. The two aerials being respectively 10 and 20 metres in height, 10 and 20 metres in length and having in each case eight wires and with a counterpoise of eight wires, firstly 10 metres long and, secondly, 20 metres long.

Abraham's theory does not apply to this case, but it is seen the resistance decreases if the antenna capacity is increased at a constant wavelength ($\lambda > 500$ metres). Also if the height of the antenna and the wavelength are kept constant a similar reduction of resistance takes place if the capacity is increased.

Fig. 8 shows these results when using a larger counterpoise. The 12-wire umbrella type aerial had, in this case, a height of 30 metres, a span of 80 metres and wire



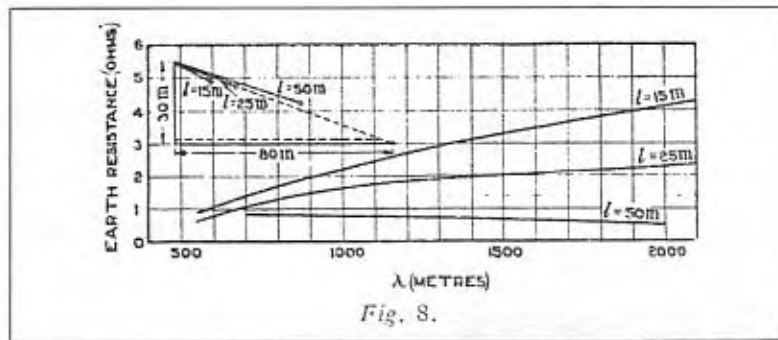


Fig. 8.

lengths of 15, 25 and 50 metres, giving capacities of 1,000, 1,400 and 2,700 centimetres respectively. The counterpoise consisted of 50 wires 40 metres in length connected with a further 100 wires 40 metres in length. The resistance was found to decrease greatly with the capacity, and could be brought down below one ohm. That is, a much lower value than was possible when using an earth connection, even if the aerial capacity was very large. Fig. 9 gives corresponding values for an L-antenna with a larger counterpoise (antenna 4 wires of 135 metres, height 21 metres, counterpoise 12 wires of 200 metres). It is seen that in this case the resistances are large for short wavelengths, but can be reduced below one ohm for long waves.

In 1915 a larger counterpoise was built at the Sayville station, in America, and measurements were made on it by Professor Zenneck. At that time the antenna was of an umbrella shape, with a radius of about 240 metres, a capacity of 11,800 centimetres and a mean height of about 6 metres. The counterpoise extended a little beyond the area covered by the antenna, and had 56 wires, each about 265 metres long at a height of 25 metres, the underground water being 6 metres deep. The total resistance of the antenna was 1.55 ohms at a wavelength of 4,640 metres, this resistance being made up of an earth resistance 0.5 to 0.6 ohm, a radiation resistance of 0.31 ohm and a coil resistance of 0.7 ohm. These counterpoises were not yet large enough, as was shown by an arrangement of Rendahl's, at Karlsborg, in Sweden. The antenna there was of the cage type, suspended between two masts 210 metres in height, the length of the cages was 400 metres and its diameter 30 metres. It consisted of 60 wires and had a capacity of 8000 centimetres. The ground below the antenna is rocky and was covered with a close counterpoise extending to a distance of more than 400 metres. The counterpoise wires were stretched at a height of 5 metres above ground at a distance of 3 metres apart, and were supported by 500 telegraph poles.

Four hundred kilometers of copper wire, one millimetre in diameter, were used for the construction of the

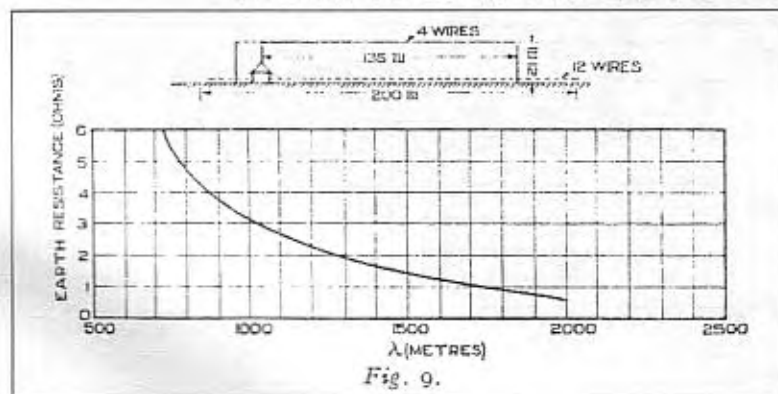


Fig. 9.

counterpoise. Rendahl gives the following values for the resistance of this antenna.

Wavelength.	Radiation Resistance.	Total Resistance.
3600	4.95	5.2
5400	2.2	2.9

As the total resistances given above include the resistance of the loading coils, and these coils, having an inductance of 90,000 centimetres, certainly possess a resistance of about 0.7 of an ohm at a wavelength of 5,400 metres, the earth resistance must be less than 0.1 of an ohm. Thus, for the first time, the ideal small earth resistance such as is needed for long-distance wireless stations has here been attained. These small values have, however, only been obtained at great cost and the expense would be even greater for a high-power station with a large area of antenna. For such antenna the counterpoise would require a space of 5 to 6 kilometres if the wires are about 2,000 metres long.

On account of the large expense, the Gesellschaft für drahtlose Telegraphie was obliged to return to earth connections. After extended experiments an arrangement was found by which similar small resistances were attained, such as those of Rendahl's using large counterpoises. The principle employed is either to make the ground connections just at the point where the lines of electric force from the antenna enter the ground. This is done so that long current paths and eddy current losses in the ground are avoided; or to distribute the current from the generator on to a number of earth connections such that each earth-point carries only as much current as corresponds to the lines of force from the antenna which enter the earth near that point.

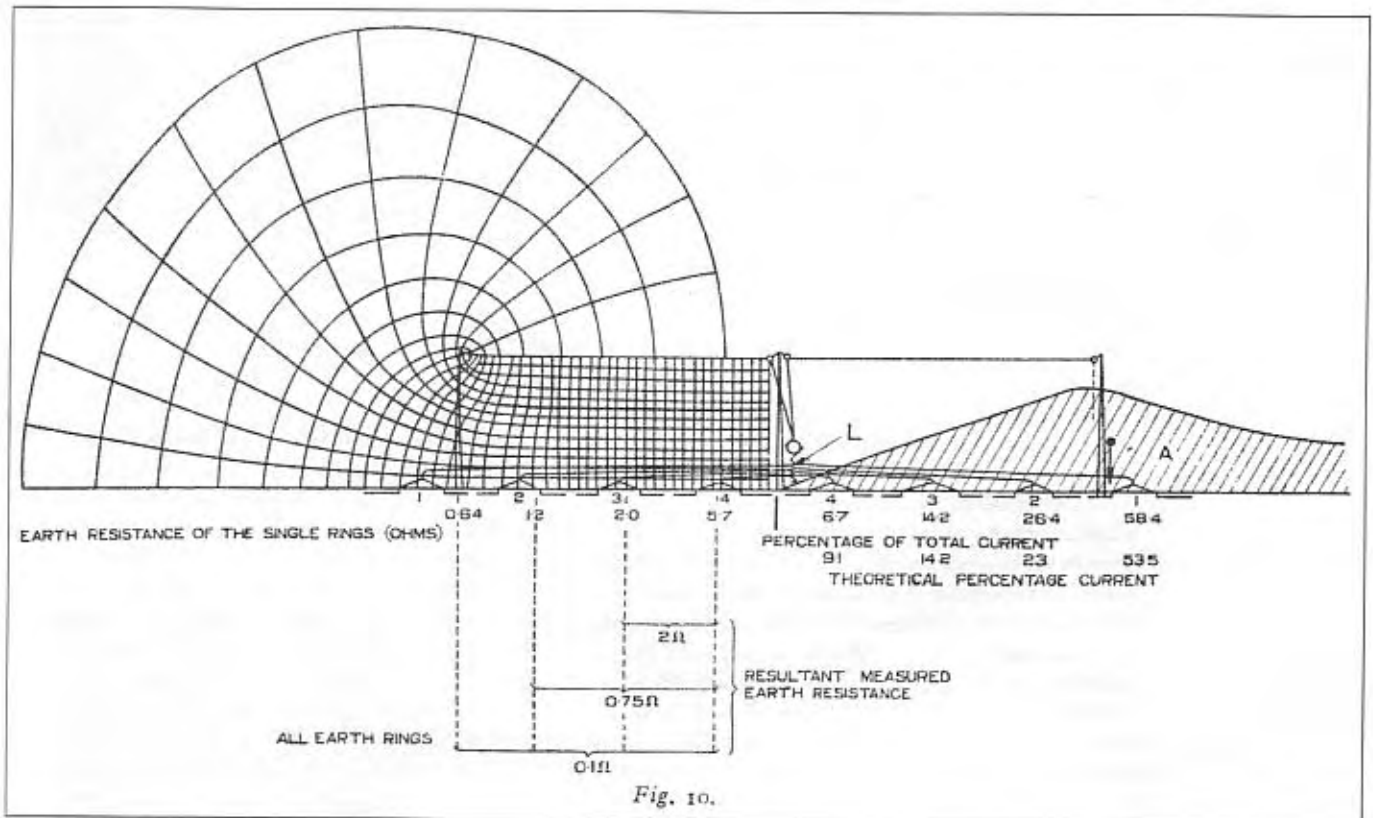
When it is not possible to cover the whole surface of the ground with earth connections, then these should be placed where the greatest number of lines of force from the antenna enter the earth.

This is where the greatest number of mistakes have been made in the design of aerial systems in the past. As usually the earth connections have been arranged in such a way that the current entered the ground at the centre of the antenna, or at a distance of about 100 metres from the centre, although the radius of the area covered by the antenna was from 400 to 500 metres. For all types of antenna the centre line of the group of lines of force between the antenna and the ground lies at the extreme border of the area covered by the antenna. This fact may also be deduced from the simple formula given by Austin for the approximate predetermination of the capacity of antenna, viz.

$$C = (9.88 a/h + \text{Sqr root } a) \times 10^5 \text{ microfarad.}$$

where a equals area of antenna and h equals its height. The capacity may be regarded as consisting of two parts. The first part is that of a simple plate condenser, having surfaces approximately equal in area to that of the antenna. The second part corresponds to the border line radiation field in the space beyond the area of the antenna. For the antennae of high-power stations having a radii of 500 metres and a height of 200 metres, the second part is larger than the first, in the ratio of 35.5×10^5 to 34.5×10^5 microfarads.

In other words, the major portion of the capacity of the whole aerial is due to the lines of force, which reach the ground beyond the area covered by the antenna. The distribution of the current entering the ground can be



obtained accurately from the field of force of the antenna, as determined by Maxwell (Fig 10).

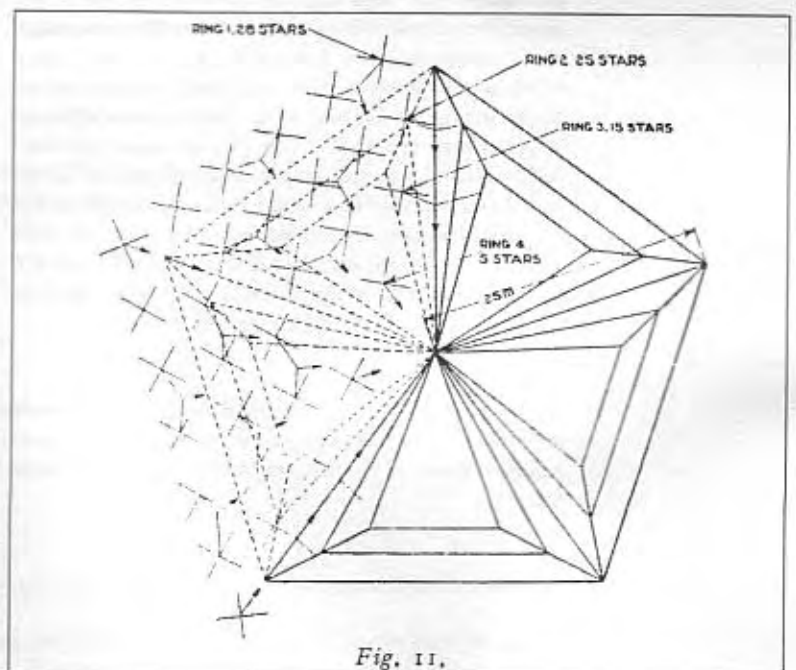
If the surface of the ground under the antenna is divided up into a number of rings of equal width (for example, one metre wide), and if the intensity of the lines of force entering each ring is determined and set out on a diagram, the total current due to the lines of force entering the ground may be represented by the shaded area A, in Fig 10

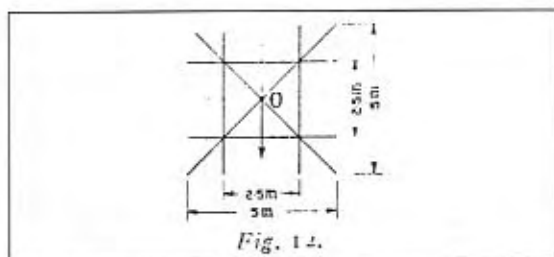
The centre of gravity of this current area lies near the border of the area covered by the antenna, so that if the current paths through the ground are to be reduced as much as possible, the earth connections should be made on the border of the antenna area and the current leads from the transmitter brought to this position.

If, however, the current is led into the ground at the centre of the antenna, the whole of the current which enters the ground beyond the area covered by the antenna (i.e., more than half of the total antenna current) must flow back through the earth along the whole length of the antenna. In the right-hand side of Fig 10 is given the current distribution of an actual antenna, which was built as a model for an antenna of a high-power station on a scale of 1:20. This antenna was pentagonal in shape, as may be seen from the plan in Fig 11. Its radius was 25 metres, its height 10 metres above the ground, and capacity 2,065 centimetres. The earth connections were arranged in four concentric rings under the whole of the antenna, and extended a distance beyond the area of the antenna equal to its height. Each earth connection was star-shaped (Fig. 12). Four conductors were led above the surface of the ground from the centre O (Fig 12) of each earthing point to the four corner earth connections. Twenty-eight earthing stars were arranged in the first ring from the outside, 25 in the second, 15 in the third, and 5 in the fourth.

Eighteen connections were made from the first ring to the centre of the aerial, 13 from the second, 8 from the third and 2 from the fourth. These connections were

secured to the central mast at a height of about 3 metres. All the earth connections in each ring were connected together, and the current distribution to the various rings of earth points could be varied as required by adjusting the connections to the coil L (Fig 10) or by adding, if necessary, more or less inductance in each earth connection. If the currents were adjusted by means of this inductance so that the whole system had the lowest resistance, the distribution of the total current in the four rings was as shown on the right-hand side of Fig 10, the outermost ring carrying more than 50 per cent. of the total current. The current distribution in each ring as calculated from the diagram of the lines of force given in the upper left-hand half of Fig 10 is set out in the second row of figures





beneath the right-hand half of Fig 10, the upper row of figures giving the actual currents that were obtained.

On the lower left-hand side of Fig 10 the earth resistances that were measured when single rings of earth connections were used are set out, the second and third lines giving the resistances when two and three rings of connections were used, and the fourth line the resistance of the whole combination of connections. When only the fourth, or inner, ring of earth connections is used, which gives an earthing scheme corresponding to those normally used, the resistance was 5.7 ohms, which is reasonable value for an earthed antenna of this size. If the outer ring only was used, the resistance fell to 0.64 ohm, but by combining all the rings together the resistance fell to less than the tenth of an ohm. This resistance was measured at various wavelengths down to 400 metres, above 1,200 metres it rises a little, and at 1,835 metres it was still less than 0.4 ohm. These measurements show that the earthings under the centre part of the antenna are of little value and could be omitted were they not required as connections to the outer border.

In later experiments the star-shaped earth connections were replaced by single pipes, by which means the earthing scheme was greatly simplified. Pipes, three centimetres in diameter and one metre in length, were driven into the ground to different depths, and the resistance measured in each case. These experiments indicated that the star-shaped connections shown in Fig 12 can be replaced by four pipes at a depth of 0.5 metre at the four corners of the star.

It has been concluded from these measurements, by consideration of Abraham's model aerial theory, that we can with certainty expect that the resistance of an antenna twenty times as large will remain far below the values quoted above, and consequently that possibly fewer rings of earth connections will be required. An earthing scheme of this type is proposed for the projected extension of the Nauen station. As the antenna will have a mean effective height of from 190 to 200 metres, the radiation resistance will be from 0.3 to 0.33 ohm at a wavelength of 12,600 metres. As the resistance of the loading coils and of the antenna wires can be reduced below 0.3 ohm, the total resistance of the antenna should lie between 0.6 and 0.7 ohm. The efficiency of the new antenna will thus be about 50 per cent, instead of 7.5 per cent., as at present.

Note

[1] This term is used throughout the article to represent the ratio $\lambda/4h$, *ie* it is a measure of the amount the wavelength of the aerial increased (by loading) above the value $4h$.



Appendix 2

The UK Amateur Allocation at 73kHz

In 1990 the Radio Society of Great Britain took an initiative to propose that there should be an LF/VLF allocation for Radio Amateurs in the UK. Since then the RSGB and the Radiocommunications Agency (RA), who licence radio communications in the UK, had been discussing the form of such an allocation. As a result a thin slice of spectrum 2.8kHz wide was found. This slot is known by UK radio amateurs as '73kHz' and was available by special permit called a 'Notice of Variation' (NoV), which is available to holders of full Class 'A' Radio Amateur licences only. An NoV was issued by the RA, in response to a request to the RSGB stating proposed activity. The technical parameters of the 73kHz Licence are as follows:

Frequency range; 71.6 to 74.4kHz

ERP (effective radiated power); 0dBW (one Watt)

Transmission modes; all those permitted for HF operation

The 73kHz allocation was to be withdrawn but has now been extended until June 2003 although no new NoVs have been issued.

The UK Amateur Allocation at 136kHz

On 30 January 1998, the 136kHz allocation was made available in the UK. The technical parameters of the 136kHz allocation are as follows:

Frequency range; 135.7kHz to 137.8kHz

ERP (effective radiated power); 0dBW (one Watt)

Transmission modes; all those permitted for HF operation

Most countries that now permit operation on this band do so within the parameters described above.

Webpages

The RSGB web site at www.rsgb.org often contains the latest news regarding LF operation.

Additions and errata to this book, including web sites and links to software will be posted at www.rsgb.org/books/extra/lf.htm.

Many LF operators have web pages, which contain a large quantity of information regarding LF. Several of these are listed below. However bear in mind the ephemeral nature of internet information and that the sites listed below were only correct when this book went to print.

G3YXM	www.picks.force9.co.uk
G0MRF	www.g0mrf.freemove.co.uk
GW4ALG	www.alg.demon.co.uk/radio/136/home.htm
G3YMC	ourworld.compuserve.com/homepages/sergeantd/136.htm
G3XDV	www.lf.thersgb.net
ON7YD	www.qsl.net/on7yd/
DL4YHF	www.qsl.net/dl4yhf
DK8KW	www.qru.de
DF8ZR	go.to/df8zr
DF3LP	www.qsl.net/df3lp
OK1FIG	www.mujweb.cz/WWW/OK1FIG/cottage.htm
I5TGC	www.dadacasa.com/i5tgc
SM6LKM	home4.swipnet.se/~w-41522/index.html
CT1DRP	homepage.esoterica.pt/~brian

rsgb_lf_group

This is an e-mail reflector which sends daily e-mails of the very latest activity and technical discussion to all who subscribe (free).

To join, simply send an e-mail to: majordomo@blacksheep.org with no subject and only **subscribe rsgb_lf_group** in the body of the message.

FFT Software – see Chapter 5**Spectrogram**

This was the first high resolution FFT software using Windows 95 and above, in conjunction with the soundcard to give LF operators the ability to decode signals many dB below the noise level. It is written by RS Horne and was originally designed to analyse birdsong. Examples of the use of versions 4.2.6.5 and 5.1.6 can be seen in Chapter 5. Spectrogram was also used in the first transatlantic crossband QSO by VE1ZJ.

Spectrogram is available at www.monumental.com/rshorne/gram.htm.

Spectran

Designed by A di Bene (I2PHD) and V De Tomasi (IK2CZL), this software is designed specifically for LF and EME operators. It is described in Chapter 5 and is going through a series of upgrades at the time of writing. Fig 5.13 shows a display from Spectran beta 2b and Fig 5.15 shows Spectran 2c.

The latest information and downloadable software is available from www.weaksignals.com.

Spectrum Lab

This is a new addition to software available for LF operators. It is written by Wolf, DL4YHF, the LF operator at DF0WD. It is now available at www.qsl.net/dl4yhf, and includes the full source code in C++Builder for Win95. Because it has only recently been released it has not been described in Chapter 5, so a brief description of its features is shown below:

- The program runs well on the latest PC or the early 90MHz Pentium.
- Contrast and Brightness control for the waterfall display. The waterfall may be repainted while the analysis is running. The waterfall and spectrum may be rotated by 90 degrees
- The frequency scale may be mirrored for LSB receivers.
- The waterfall may be scrolled back without stopping the analysis.
- The colour palette controls are useful for extracting very weak signals out of the noise .
- The waterfall scroll rate can be adjusted independent of the FFT sample size.
- In addition any combination of lowpass, highpass, bandpass or notch filter can be designed and saved to a filter file.

EasyGram

This is another program for extracting signals from the noise and was designed by Petr Maly, OK1FIG. This software is in fact a user interface for SPECTRUM.DLL (the heart of Spectrogram by RS Horne). It provides the same functions as Spectrogram, but it has a more user-friendly access. The main differences are:

- All the settings can be done while scanning is running (like Spectran).
- The picture moves in a separate window. This enables changes to be made to the size and position of the window by entering numbers in the configuration. Also, the scanning window can be maximized across the whole screen.
- All the settings can be saved to a 'profile'. Once the profile is defined, you can select all the settings at once by selecting the profile item in a combo box. This also includes position and size of the scanning window.
- Vertical (waterfall) scanning is also available.
- When you click in the scanning window, a small popup window is presented allowing either zoom to the selected freq (signal trace), or setting a marker to be selected.
- Saving to .BMP. A unique file name is generated from system's date/time, or the user can change it.
- Saving to files can be triggered by a timer. The timing intervals from 5 seconds to 60 minutes can be selected.
- A simple browser is included so that many bitmap files can be viewed.
- Even more settings are saved to the INI file. Some of them are set to the selected profile, some of them are global.

The program can be downloaded from: mujweb.cz/www/ok1fig/EasyGram.htm.

Inductance Calculation Program

A program for calculating loading coil inductance by Reg Edwards, G4FGQ, is available at: www.btinternet.com/~g4fgq.regp.

Suppliers, Electronic Components

Farnell Electronic Components, Canal Road, Leeds LS12 2TU.

Tel: +44 (0)1132 636311. Website: <http://www.farnell.com>

Arrow Tel: +44 (0)1279 626777

Bonex Tel: +44 (0)1753 549502

Mainline Tel: +44 (0)1162 777648

Suppliers, Antenna Materials

W. H. Westlake, Clawton, Holsworthy, Devon EX22 6QN.

Tel: +44 (0)1409 253758

Sending Slow Morse

Chapter 5 describes the facilities of these programs.

SLOWCW: See www.g3wkl.freemove.co.uk/lf/files.htm

QRS: See www.qsl.net/on7yd/136khz.htm



Appendix 3

Crossing the Atlantic on LF

Just as this book was going to press, 10 September 2000, this e-mail appeared on the *rsgb_lf_group* e-mail reflector:

A crossband QSO between the UK and Canada was completed at 0008Z on 10 Sept 2000. G0MRF was operating on 135.71kHz and VE1ZJ was operating on 14,043kHz details w/ follow tmw. Tnx for great effort David. 73 de John VE1ZJ

This historic QSO was the result of some close liaison with all concerned. It is remarkable because it followed tuition sessions by e-mail and phone by Dave to John (a very experienced operator on HF and VHF EME but a novice on LF) on the use of LF receiving techniques and the use of Spectrogram. It is surprising what can be achieved with the right person, in the right place at the right time, who is willing to listen.

David G0MRF describes his side of the operation as follows:

My operation on 9/10 Sept was a continuation of an attempt to construct and test a loaded dipole for 136kHz. Despite building a scale model at 2.5MHz, Saturday night's experiment failed. I managed to get a reasonable match but discovered that all the current was flowing around the link coupling coil with none actually making it to the elements. Oh well!

The location for G0MRF was a West London tower block in Feltham, just South of Heathrow airport. The 15th story, top floor flat, is the home of Sean Griffin 2E1AXK. We were joined later by Tony Fell G7DGW. The dipole was 're-worked' to give a standard sloping antenna with two wires each 250 feet long. The angle between the two was about 60 – 80 degrees. The ends of the wires were about 30 feet above ground and tied off with 50lb breaking-strain fishing line. The earth was connected to the plumbing system of the building. The antenna was of course fed from the top, at about 160 feet above ground.

Loading was a combination of fixed inductor in series with a variable coil. Matching was achieved by earthing the base of the fixed coil and tapping a few turns up.

Once this was wired up I used an amplified DDS source to drive a modified version of the reflectometer bridge published in *Radcom* 9/2000. With this little circuit (3 x 50R resistors and a transformer) the antenna loading and

matching was completed in less than 3 minutes.

The transmitter was a converted Decca unit with a DC input of 58 volts at 15 amps. Antenna current peaked at 7.2 amps. Drive was a 'SM6LKM' DDS with 0.25Hz steps at 136k.

John VE1ZJ, Larry VA3LK and Jack VE1ZZ on the other side of the Atlantic were all aware of the schedule and frequency for the tests. After about an hour John called to confirm he could see the signals and would I QSY up 1Hz to clear a Loran line. No problem.

The crossband QSO was then completed without problem with 'O' signals on both 135.711 and 14.043 MHz.

Interestingly I then reduced power to just 400W DC input, or about 320W Tx output and John reported an M signal. Even allowing for the large antenna I believe this shows that many UK/EU stations will be able to make the transatlantic path this winter.

In another telephone conversation, John reported that Jack VE1ZZ was transmitting on 136.5kHz using a long horizontal antenna (OH1TN/G3OLB style). Unfortunately, Jack does not have a computer for QRS. Using Spectran with about 40Hz across the screen we looked very closely for Jack's 10 second dashes. After about a minute we spotted several dashes, broken by static, about 8Hz above 136.500. In a phone call to Jack I asked are you transmitting now 8Hz off frequency? "No" said Jack, "my counter says its only 7Hz"! I decided I could settle for that.

In one evening: 136kHz signals from the UK heard in Canada; a QSO crossband to 20m G/VE; 136kHz signals transmitted from Canada heard in the UK.

John VE1ZJ describes the contact from his end:

My summer QTH (locator FN95) is where the radio activity took place. The town is Big Pond, Nova Scotia (formerly known as Loch Mhor). The Bras d'Or lakes, an inland sea is 30 feet from my antenna 32ft vertical with one radial in the saltwater. The Atlantic is approximately 20 miles to the east.

The sequence of the crossband QSO was as follows: Dave had a surprisingly strong signal in to Cape Breton Island, Nova Scotia Canada. His frequency was 135.711kHz.

Weak dashes were first observed at just after 2205Z (2218 my sunset). The German commercial LF station on 128.9kHz was S-3 and the one on 138.9kHz was S-2. The noise was extremely low.

At 2233UTC, 128.9 = S-4.5 and 138.9 = S-1. Lot of

dashes. It looked like G0MRF was coming across the pond. At 2240UTC, 128.9 = S-5 and 128.9 = S-3.

Finally, at 2245UTC had solid copy on G0MRF, with the crossband QSO 135.71kHz to 14.043MHz completed 10 Sept 0008Z

Dave's signals were 'O' copy. I could see every dot and dash at 0000UTC on the 10th. At that time 128.9 = S-6 and 138.9 = S-6. S-7 represents field strength of 100 microvolts per metre. I would estimate Dave's signals at a little less than S-0, at least 36dB below DCF39 (138.9kHz).

After 0100 the signals dropped off to 'M' copy and by 0250UTC they were down to a 'T' and fading. I did not see them on Spectrogram again. I believe the signal-to-noise ratio dropped because QRN went up. In addition I think we had QRM from the Marathon station.

The QRM had a spectral line at 135.782kHz and began to fade by 0245 to 'M' copy, 128.9 = S -8.5 and 138.9 = S5.5. The noise, however, was now about 4 S-units higher than at sunset. I did not see David after 0250 on 135.71.

I am amazed that in EU a remote transmitter site can set up in only two or three hours with such good results. Congratulations and thanks all for helping me with receiving tips.



The signal from G0MRF, as seen by VE1ZJ, using Spectrogram. The parallel lines are sidebands of the Loran navigation system.

Further activity:

This achievement triggered off a lot of activity and several LF operators transmitted in an informally organised group at the bottom end of the band between 137.705kHz and 135.55kHz. This resulted in some successes, including G3LDO and M0BMU who used 'normal' backyard antennas to reach VE1ZJ on LF, though ironically the 14MHz reverse link failed. A second crossband contact took place, this time between VE1ZJ and OK1FIG at 2245UTC on 24 September, 2000. OK1FIG had the use of a 150m mast for the weekend!



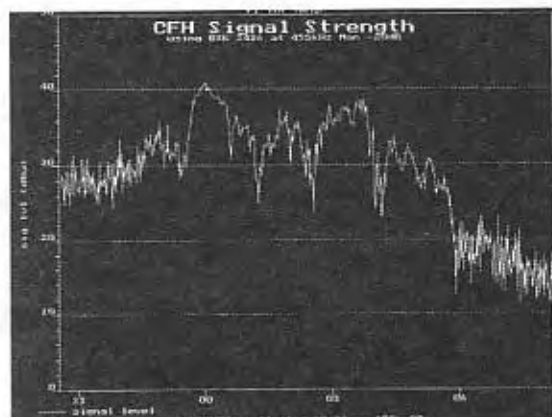
The transmissions on the night of the 23 of September, 2000, as seen by Dutch SWL, Ko Versteeg, NL9222.



VE1ZJ also received signals from M0BMU (top trace: the letters BMU can be seen in the middle) and G3LDO (bottom trace: the letters DO, then G followed by the two dashes of the '3', then the di-dah-dit of the 'L').

Researching suitable propagation:

Alan G3NYK has been monitoring CFH, an RTTY station on 137.0kHz, located in Nova Scotia. By recording the signal strength every night and making the data available on the LF reflector, it gives the test group an insight into transatlantic conditions during the course of the night.



This is a plot of the calibrated signal received at G3NYK on the night that G3LDO was received in Big Pond NS. The significant part seems to be the peak to 40dBu, which is the time that Peter's signal was seen (all times are UTC).

Alan describes his monitoring system as follows: The logging software is PicoLog from Pico Technology. The 455kHz output from the AOR7030 fed to a B&K model 2426 autoranging mV meter. The autoranging isn't used, but averaging and time-constant (set to slow ~2.7 secs) and pseudo log scale are useful. The DC output from the meter is sampled by the PicoTech ADC. The whole system is calibrated by putting a level into the AOR from a calibrated signal generator (Marconi TF202B) and reading off the millivolts recorded by the ADC. Then a calibration table is entered into the Pico software to enable it to scale (piecewise linear interpolation) the RX input signal directly in dBu. The system starts to saturate at about 300µV when I run out of headroom in the B&K 2426. The ideal would be a proper logarithmic amp after the AOR. This works and is useful (and easy) – the main thing is to collect a lot of data if I am to understand what is happening. I have never actually measured signal level before. I found using the AGC on the AOR not ideal and prone to nearby strong signals.



HIP4080

80V/2.5A Peak, High Frequency Full Bridge FET Driver

March 1995

Features

- Drives N-Channel FET Full Bridge Including High Side Chop Capability
- Bootstrap Supply Max Voltage to 95VDC
- Drives 1000pF Load at 1MHz in Free Air at +50°C with Rise and Fall Times of Typically 10ns
- User-Programmable Dead Time
- Charge-Pump and Bootstrap Maintain Upper Bias Supplies
- DIS (Disable) Pin Pulls Gates Low
- Input Logic Thresholds Compatible with 5V to 15V Logic Levels
- Very Low Power Consumption

Applications

- Medium/Large Voice Coil Motors
- Full Bridge Power Supplies
- Class D Audio Power Amplifiers
- High Performance Motor Controls
- Noise Cancellation Systems
- Battery Powered Vehicles
- Peripherals
- U.P.S.

Description

The HIP4080 is a high frequency, medium voltage Full Bridge N-Channel FET driver IC, available in 20 lead plastic SOIC and DIP packages. The HIP4080 includes an input comparator, used to facilitate the "hysteresis" and PWM modes of operation. Its HEN (high enable) lead can force current to freewheel in the bottom two external power MOSFETs, maintaining the upper power MOSFETs off. Since it can switch at frequencies up to 1MHz, the HIP4080 is well suited for driving Voice Coil Motors, switching amplifiers in class D high-frequency switching audio amplifiers and power supplies.

HIP4080 can also drive medium voltage brush motors, and two HIP4080s can be used to drive high performance stepper motors, since the short minimum "on-time" can provide fine micro-stepping capability.

Short propagation delays of approximately 55ns maximizes control loop crossover frequencies and dead-times which can be adjusted to near zero to minimize distortion, resulting in precise control of the driven load.

The similar HIP4081 IC allows independent control of all 4 FETs in an Full Bridge configuration.

See also, Application Note AN9324 for the HIP4080.

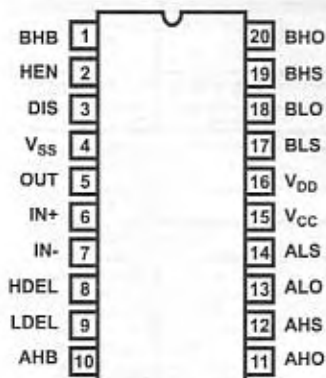
Similar part, HIP4080A, includes under voltage circuitry which doesn't require the circuitry shown in Figure 30 of this data sheet.

Ordering Information

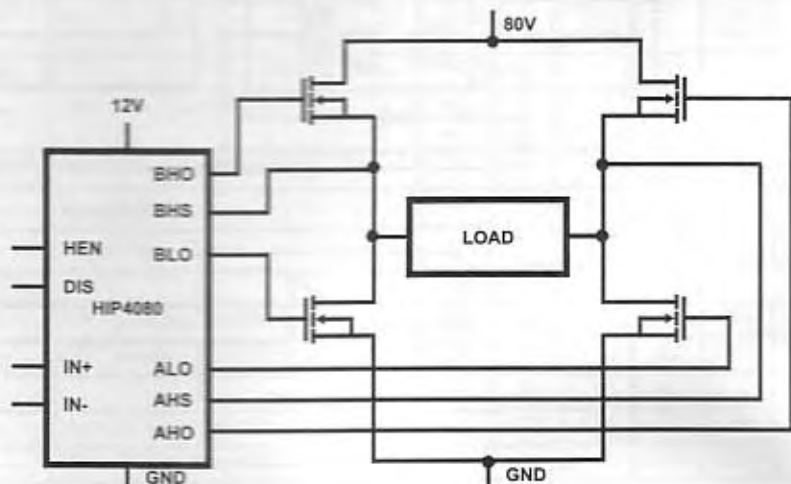
PART NUMBER	TEMPERATURE RANGE	PACKAGE
HIP4080IP	-40°C to +85°C	20 Lead Plastic DIP
HIP4080IB	-40°C to +85°C	20 Lead Plastic SOIC (W)

Pinout

HIP4080 (PDIP, SOIC)
TOP VIEW

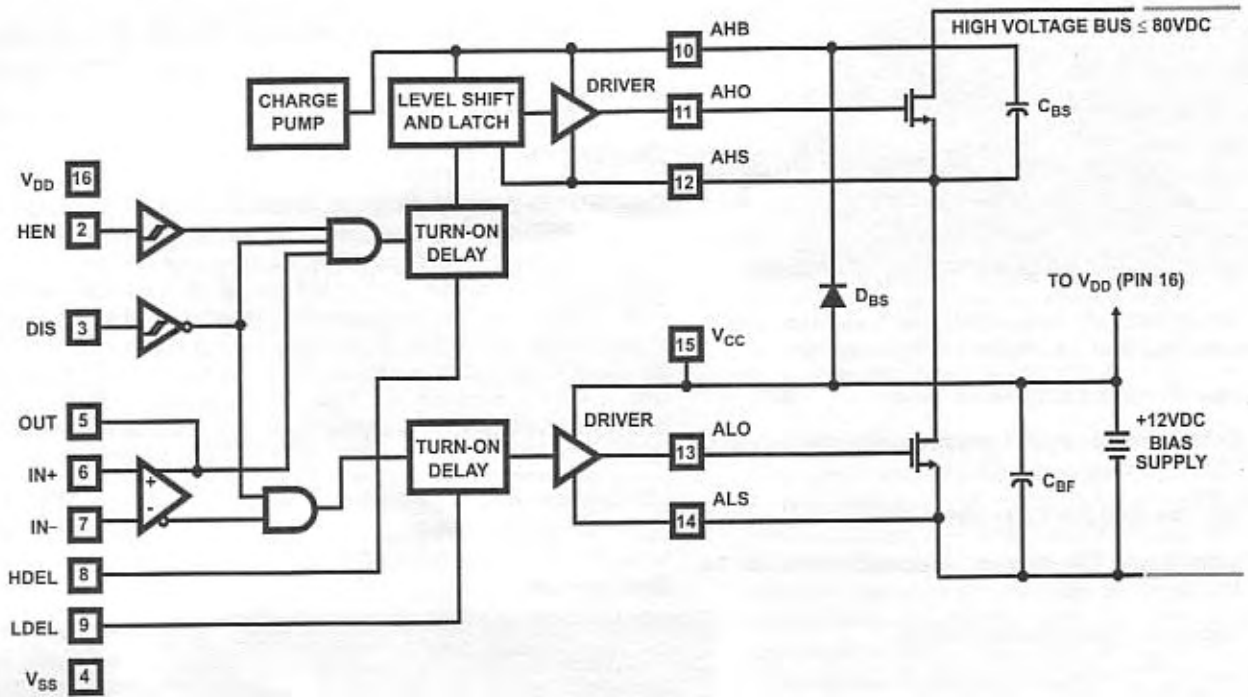


Application Block Diagram

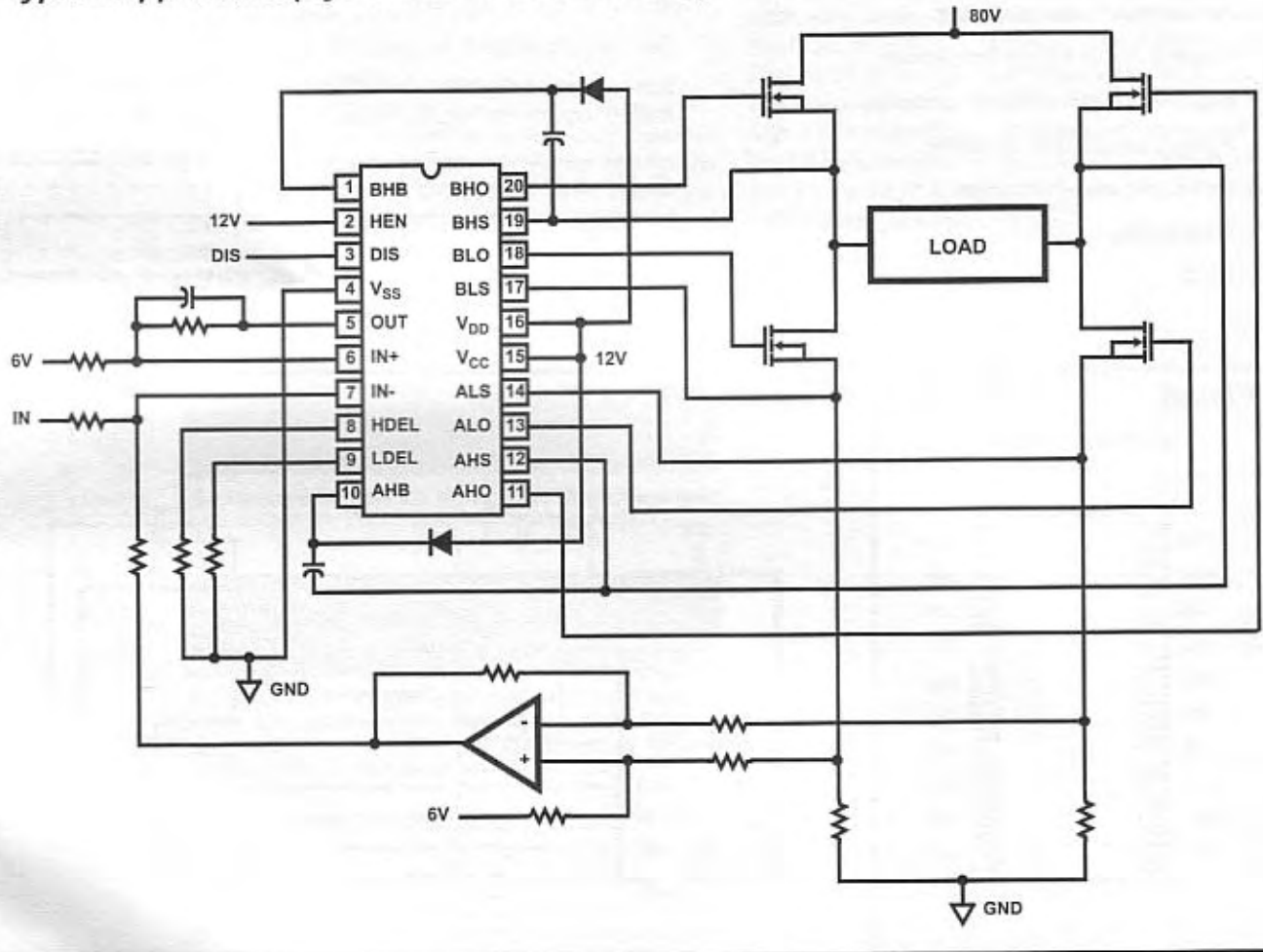


HIP4080

Functional Block Diagram (1/2 HIP4080)



Typical Application (Hysteresis Mode Switching)



Specifications HIP4080

Absolute Maximum Ratings

Supply Voltage, V_{DD} and V_{CC} -0.3V to 16V
 Logic I/O Voltages -0.3V to V_{DD} +0.3V
 Voltage on AHS, BHS -6.0V (Transient) to 80V (25°C to 125°C)
 Voltage on AHS, BHS -6.0V (Transient) to 70V (-55°C to 125°C)
 Voltage on ALS, BLS -2.0V (Transient) to +2.0V (Transient)
 Voltage on AHB, BHB $V_{AHS, BHS}$ -0.3V to $V_{AHS, BHS}$ +16V
 Voltage on ALO, BLO $V_{ALS, BLS}$ -0.3V to V_{CC} +0.3V
 Voltage on AHO, BHO $V_{AHS, BHS}$ -0.3V to $V_{AHB, BHB}$ +0.3V
 Input Current, HDEL and LDEL -5mA to 0mA
 Phase Slew Rate 20V/ns
 NOTE: All Voltages relative to pin 4, V_{SS} , unless otherwise specified.

Thermal Information

Thermal Resistance θ_{JA}
 SOIC Package +85°C/W
 DIP Package +75°C/W
 Maximum Power Dissipation at +85°C
 SOIC Package 470mW
 DIP Package 530mW
 Storage Temperature Range -65°C to +150°C
 Operating Max. Junction Temperature +125°C
 Lead Temperature (Soldering 10s) +300°C
 (For SOIC - Lead Tips Only)

CAUTION: Stresses above those listed in "Absolute Maximum Ratings" may cause permanent damage to the device. This is a stress only rating and operation of the device at these or any other conditions above those indicated in the operational sections of this specification is not implied.

Operating Conditions

Supply Voltage, V_{DD} and V_{CC} +8V to +15V
 Voltage on ALS, BLS -1.0V to +1.0V
 Voltage on AHB, BHB $V_{AHS, BHS}$ +5V to $V_{AHS, BHS}$ +15V
 Input Current, HDEL and LDEL -500 μ A to -50 μ A
 Operating Ambient Temperature Range -40°C to +85°C

Electrical Specifications $V_{DD} = V_{CC} = V_{AHB} = V_{BHB} = 12V$, $V_{SS} = V_{ALS} = V_{BLS} = V_{AHS} = V_{BHS} = 0V$, $R_{HDEL} = R_{LDEL} = 100K$, and $T_A = +25^\circ C$, Unless Otherwise Specified

PARAMETERS	SYMBOL	TEST CONDITIONS	$T_J = +25^\circ C$			$T_J = -40^\circ C$ TO +125°C		UNITS
			MIN	TYP	MAX	MIN	MAX	
SUPPLY CURRENTS AND CHARGE PUMPS								
V_{DD} Quiescent Current	I_{DD}	IN- = 2.5V, Other Inputs = 0V	8	10.5	13	7	14	mA
V_{DD} Operating Current	I_{DDO}	Outputs switching f = 500kHz	9	11	14	8	15	mA
V_{CC} Quiescent Current	I_{CC}	IN- = 2.5V, Other Inputs = 0V, $I_{ALO} = I_{BLO} = 0$	-	25	80	-	100	μ A
V_{CC} Operating Current	I_{CCO}	f = 500kHz, No Load	1	1.5	2.0	0.8	3	mA
AHB, BHB Quiescent Current - Qpump Output Current	I_{AHB}, I_{BHB}	IN- = 2.5V, Other Inputs = 0V, $I_{AHO} =$ $I_{BHO} = 0$, $V_{DD} = V_{CC} = V_{AHB} = V_{BHB} = 10V$	-50	-30	-15	-60	-10	μ A
AHB, BHB Operating Current	I_{AHBO}, I_{BHBO}	f = 500kHz, No Load	0.5	0.9	1.3	0.4	1.7	mA
AHS, BHS, AHB, BHB Leakage Current	I_{HLK}	$V_{AHS} = V_{BHS} = V_{AHB} = V_{BHB} = 95V$	-	0.02	1.0	-	10	μ A
AHB-AHS, BHB-BHS Qpump Output Voltage	$V_{AHB} - V_{AHS}$ $V_{BHB} - V_{BHS}$	$I_{AHB} = I_{AHS} = 0$, No Load	11.5	12.6	14.0	10.5	14.5	V
INPUT COMPARATOR PINS: IN+, IN-, OUT								
Offset Voltage	V_{OS}	Over Common Mode Voltage Range	-10	0	+10	-15	+15	mV
Input Bias Current	I_{IB}		0	0.5	2	0	4	μ A
Input Offset Current	I_{OS}		-1	0	+1	-2	+2	μ A
Input Common Mode Voltage Range	CMVR		1	-	V_{DD} -1.5	1	V_{DD} -1.5	V
Voltage Gain	AVOL		10	25	-	10	-	V/mV
OUT High Level Output Voltage	V_{OH}	IN+ > IN-, $I_{OH} = -300\mu A$	V_{DD} -0.4	-	-	V_{DD} -0.5	-	V
OUT Low Level Output Voltage	V_{OL}	IN+ < IN-, $I_{OL} = 300\mu A$	-	-	0.3	-	0.4	V
High Level Output Current	I_{OH}	$V_{OUT} = 6V$	-9	-7	-4	-11	-2	mA
Low Level Output Current	I_{OL}	$V_{OUT} = 6V$	8	10	12	5	14	mA
INPUT PINS: DIS								
Low Level Input Voltage	V_{IL}	Full Operating Conditions	-	-	1.0	-	0.8	V
High Level Input Voltage	V_{IH}	Full Operating Conditions	2.5	-	-	2.7	-	V
Input Voltage Hysteresis			-	35	-	-	-	mV
Low Level Input Current	I_{IL}	$V_{IN} = 0V$, Full Operating Conditions	-130	-100	-75	-135	-65	μ A
High Level Input Current	I_{IH}	$V_{IN} = 5V$, Full Operating Conditions	-1	-	+1	-10	+10	μ A

Specifications HIP4080

Electrical Specifications $V_{DD} = V_{CC} = V_{AHB} = V_{BHB} = 12V$, $V_{SS} = V_{ALS} = V_{BLS} = V_{AHS} = V_{BHS} = 0V$, $R_{HDEL} = R_{LDEL} = 100K$, and $T_A = +25^\circ C$, Unless Otherwise Specified (Continued)

PARAMETERS	SYMBOL	TEST CONDITIONS	$T_J = +25^\circ C$			$T_J = -40^\circ C$ TO $+125^\circ C$		UNITS
			MIN	TYP	MAX	MIN	MAX	
INPUT PINS: HEN								
Low Level Input Voltage	V_{IL}	Full Operating Conditions	-	-	1.0	-	0.8	V
High Level Input Voltage	V_{IH}	Full Operating Conditions	2.5	-	-	2.7	-	V
Input Voltage Hysteresis			-	35	-	-	-	mV
Low Level Input Current	I_{IL}	$V_{IN} = 0V$, Full Operating Conditions	-260	-200	-150	-270	-130	μA
High Level Input Current	I_{IH}	$V_{IN} = 5V$, Full Operating Conditions	-1	-	+1	-10	+10	μA
TURN-ON DELAY PINS: LDEL AND HDEL								
LDEL, HDEL Voltage	$V_{HDEL, V}$	$I_{HDEL} = I_{LDEL} = -100\mu A$	4.9	5.1	5.3	4.8	5.4	V
GATE DRIVER OUTPUT PINS: ALO, BLO, AHO, AND BHO								
Low Level Output Voltage	V_{OL}	$I_{OUT} = 100mA$.70	0.85	1.0	0.5	1.1	V
High Level Output Voltage	$V_{CC} - V_{OH}$	$I_{OUT} = -100mA$	0.8	0.95	1.1	0.5	1.2	V
Peak Pull-up Current	I_{O+}	$V_{OUT} = 0V$	1.7	2.6	3.8	1.4	4.1	A
Peak Pull-down Current	I_{O-}	$V_{OUT} = 12V$	1.7	2.4	3.3	1.3	3.6	A

Switching Specifications $V_{DD} = V_{CC} = V_{AHB} = V_{BHB} = 12V$, $V_{SS} = V_{ALS} = V_{BLS} = V_{AHS} = V_{BHS} = 0V$, $R_{HDEL} = R_{LDEL} = 10K$, $C_L = 1000pF$, and $T_A = +25^\circ C$, Unless Otherwise Specified

PARAMETERS	SYMBOL	TEST CONDITIONS	$T_J = +25^\circ C$			$T_J = -40^\circ C$ TO $+125^\circ C$		UNITS
			MIN	TYP	MAX	MIN	MAX	
Lower Turn-off Propagation Delay (IN+/IN- to ALO/BLO)	T_{LPHL}		-	40	70	-	90	ns
Upper Turn-off Propagation Delay (IN+/IN- to AHO/BHO)	T_{HPHL}		-	50	80	-	110	ns
Lower Turn-on Propagation Delay (IN+/IN- to ALO/BLO)	T_{LPLH}	$R_{HDEL} = R_{LDEL} = 10K$	-	45	70	-	90	ns
Upper Turn-on Propagation Delay (IN+/IN- to AHO/BHO)	T_{HPLH}	$R_{HDEL} = R_{LDEL} = 10K$	-	70	110	-	140	ns
Rise Time	T_R		-	10	25	-	35	ns
Fall Time	T_F		-	10	25	-	35	ns
Turn-on Input Pulse Width	$T_{PWIN-ON}$	$R_{HDEL} = R_{LDEL} = 10K$	50	-	-	50	-	ns
Turn-off Input Pulse Width	$T_{PWIN-OFF}$	$R_{HDEL} = R_{LDEL} = 10K$	40	-	-	40	-	ns
Disable Turn-off Propagation Delay (DIS - Lower Outputs)	T_{DISLOW}		-	45	75	-	95	ns
Disable Turn-off Propagation Delay (DIS - Upper Outputs)	$T_{DISHIGH}$		-	55	85	-	105	ns
Disable to Lower Turn-on Propagation Delay (DIS - ALO and BLO)	T_{DLPLH}		-	35	70	-	90	ns
Refresh Pulse Width (ALO and BLO)	T_{REF-PW}		160	260	380	140	420	ns
Disable to Upper Enable (DIS - AHO and BHO)	T_{UEN}		-	335	500	-	550	ns
HEN-AHO, BHO Turn-off, Propagation Delay	$T_{HEN-PHL}$	$R_{HDEL} = R_{LDEL} = 10K$	-	35	70	-	90	ns
HEN-AHO, BHO Turn-on, Propagation Delay	$T_{HEN-PLH}$	$R_{HDEL} = R_{LDEL} = 10K$	-	60	90	-	110	ns

TRUTH TABLE

INPUT			OUTPUT			
IN+ > IN-	HEN	DIS	ALO	AHO	BLO	BHO
X	X	1	0	0	0	0
1	1	0	0	1	1	0
0	1	0	1	0	0	1
1	0	0	0	0	1	0
0	0	0	1	0	0	0

HIP4080

Pin Descriptions

PIN NUMBER	SYMBOL	DESCRIPTION
1	BHB	B High-side Bootstrap supply. External bootstrap diode and capacitor are required. Connect cathode of bootstrap diode and positive side of bootstrap capacitor to this pin. Internal charge pump supplies 30 μ A out of this pin to maintain bootstrap supply. Internal circuitry clamps the bootstrap supply to approximately 12.8V.
2	HEN	High-side Enable input. Logic level input that when low overrides IN+/IN- (Pins 6 and 7) to put AHO and BHO drivers (Pins 11 and 20) in low output state. When HEN is high AHO and BHO are controlled by IN+/IN- inputs. The pin can be driven by signal levels of 0V to 15V (no greater than V_{DD}). An internal 100 μ A pull-up to V_{DD} will hold HEN high, so no connection is required if high-side and low-side outputs are to be controlled by IN+/IN- inputs.
3	DIS	DISable input. Logic level input that when taken high sets all four outputs low. DIS high overrides all other inputs. When DIS is taken low the outputs are controlled by the other inputs. The pin can be driven by signal levels of 0V to 15V (no greater than V_{DD}). An internal 100 μ A pull-up to V_{DD} will hold DIS high if this pin is not driven.
4	V_{SS}	Chip negative supply, generally will be ground.
5	OUT	OUTput of the input control comparator. This output can be used for feedback and hysteresis.
6	IN+	Non-inverting input of control comparator. If IN+ is greater than IN- (Pin 7) then ALO and BHO are low level outputs and BLO and AHO are high level outputs. If IN+ is less than IN- then ALO and BHO are high level outputs and BLO and AHO are low level outputs. DIS (Pin 3) high level will override IN+/IN- control for all outputs. HEN (Pin 2) low level will override IN+/IN- control of AHO and BHO. When switching in four quadrant mode, dead time in a half bridge leg is controlled by HDEL and LDEL (Pins 8 and 9).
7	IN-	Inverting input of control comparator. See IN+ (Pin 6) description.
8	HDEL	High-side turn-on DELay. Connect resistor from this pin to V_{SS} to set timing current that defines the turn-on delay of both high-side drivers. The low-side drivers turn-off with no adjustable delay, so the HDEL resistor guarantees no shoot-through by delaying the turn-on of the high-side drivers. HDEL reference voltage is approximately 5.1V.
9	LDEL	Low-side turn-on DELay. Connect resistor from this pin to V_{SS} to set timing current that defines the turn-on delay of both low-side drivers. The high-side drivers turn-off with no adjustable delay, so the LDEL resistor guarantees no shoot-through by delaying the turn-on of the low-side drivers. LDEL reference voltage is approximately 5.1V.
10	AHB	A High-side Bootstrap supply. External bootstrap diode and capacitor are required. Connect cathode of bootstrap diode and positive side of bootstrap capacitor to this pin. Internal charge pump supplies 30 μ A out of this pin to maintain bootstrap supply. Internal circuitry clamps the bootstrap supply to approximately 12.8V.
11	AHO	A High-side Output. Connect to gate of A High-side power MOSFET.
12	AHS	A High-side Source connection. Connect to source of A High-side power MOSFET. Connect negative side of bootstrap capacitor to this pin.
13	ALO	A Low-side Output. Connect to gate of A Low-side power MOSFET.
14	ALS	A Low-side Source connection. Connect to source of A Low-side power MOSFET.
15	V_{CC}	Positive supply to gate drivers. Must be same potential as V_{DD} (Pin 16). Connect to anodes of two bootstrap diodes.
16	V_{DD}	Positive supply to lower gate drivers. Must be same potential as V_{CC} (Pin 15). De-couple this pin to V_{SS} (Pin 4).
17	BLS	B Low-side Source connection. Connect to source of B Low-side power MOSFET.
18	BLO	B Low-side Output. Connect to gate of B Low-side power MOSFET.
19	BHS	B High-side Source connection. Connect to source of B High-side power MOSFET. Connect negative side of bootstrap capacitor to this pin.
20	BHO	B High-side Output. Connect to gate of B High-side power MOSFET.

HIP4080

Timing Diagrams

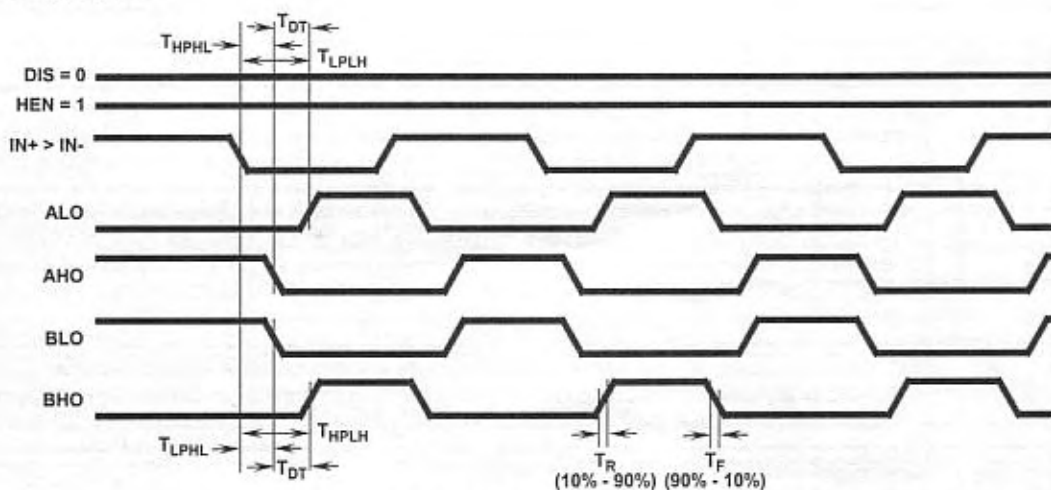


FIGURE 1. BISTATE MODE

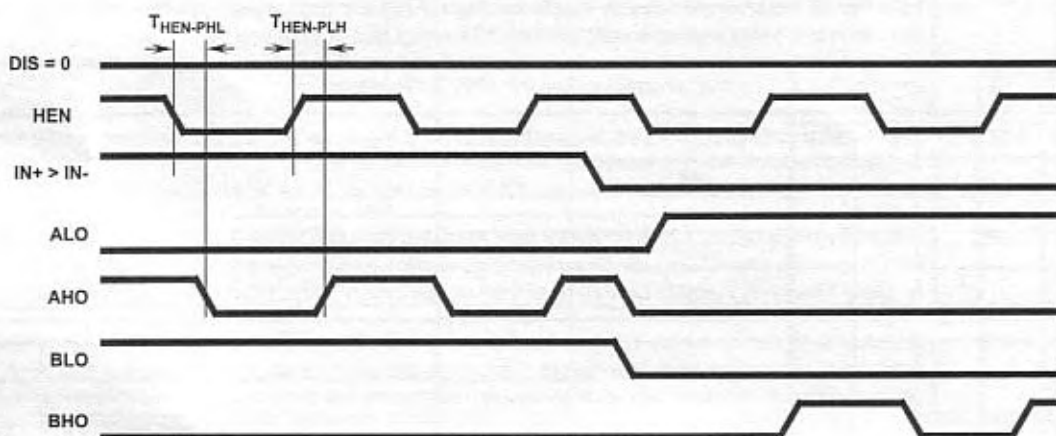


FIGURE 2. HIGH SIDE CHOP MODE

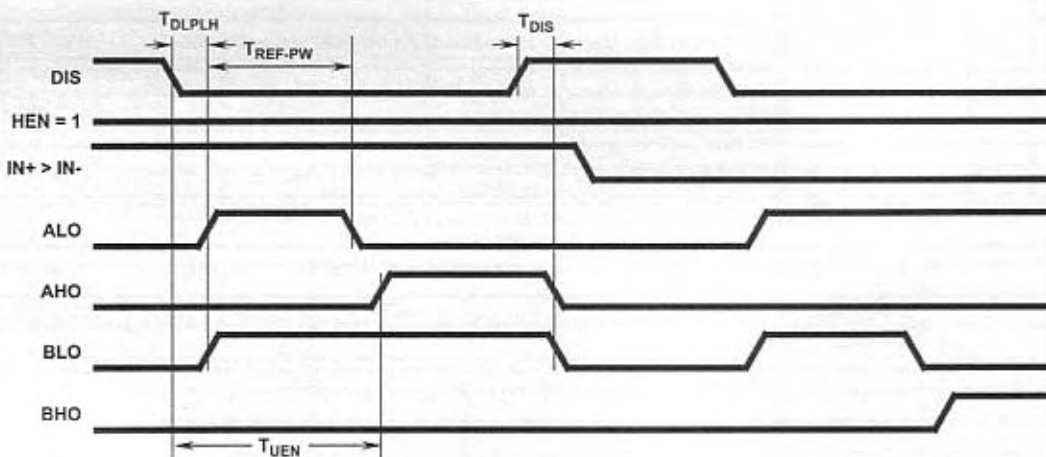


FIGURE 3. DISABLE FUNCTION

HIP4080

Typical Performance Curves $V_{DD} = V_{CC} = V_{AHB} = V_{BHB} = 12V$, $V_{SS} = V_{ALS} = V_{BLS} = V_{AHS} = V_{BHS} = 0V$,
 $R_{HDEL} = R_{LDEL} = 100K$, and $T_A = +25^\circ C$, Unless Otherwise Specified

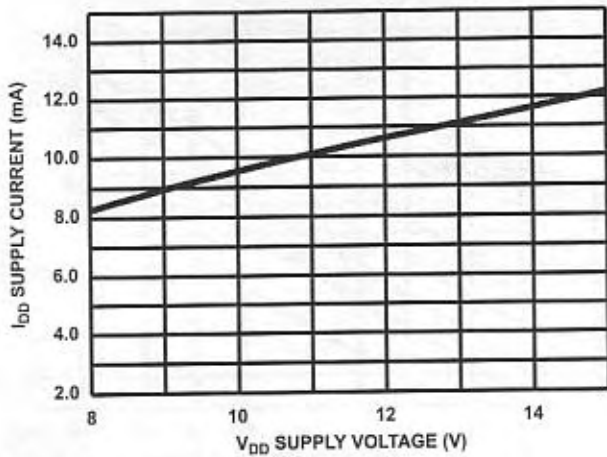


FIGURE 4. QUIESCENT I_{DD} SUPPLY CURRENT vs V_{DD} SUPPLY VOLTAGE

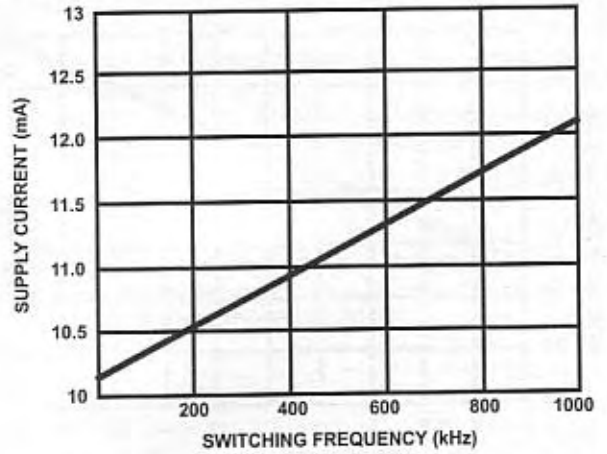


FIGURE 5. I_{DD0} , NO-LOAD I_{DD} SUPPLY CURRENT vs FREQUENCY (kHz)

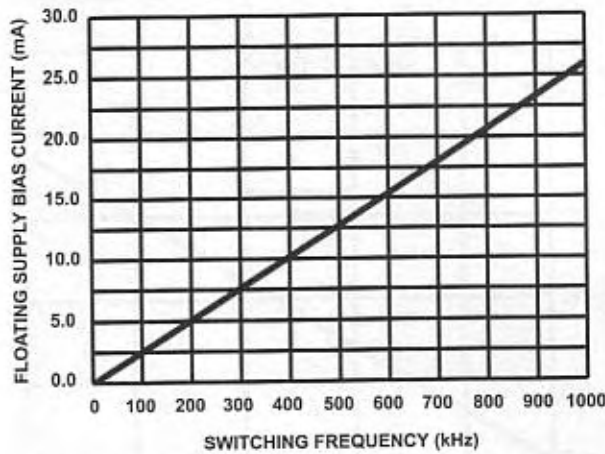


FIGURE 6. SIDE A, B FLOATING SUPPLY BIAS CURRENT vs FREQUENCY (LOAD = 1000pF)

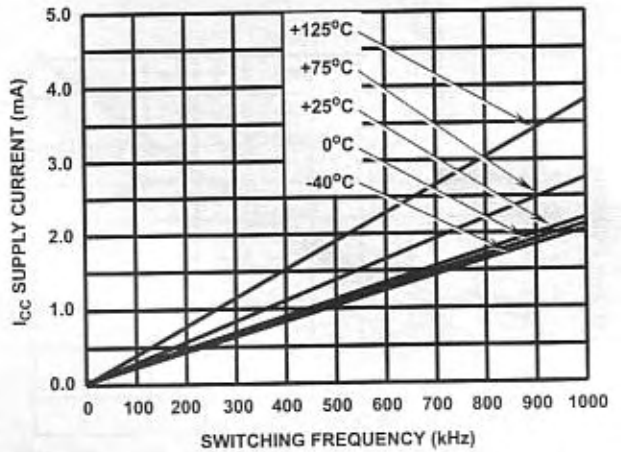


FIGURE 7. I_{CC0} , NO-LOAD I_{CC} SUPPLY CURRENT vs FREQUENCY (kHz) TEMPERATURE

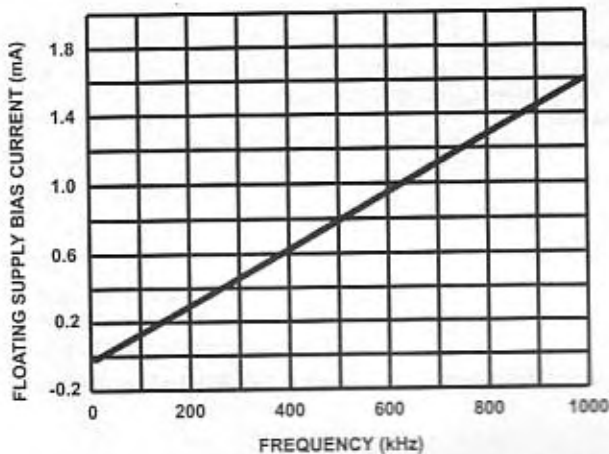


FIGURE 8. I_{AHB} , I_{BHB} , NO-LOAD FLOATING SUPPLY BIAS CURRENT vs FREQUENCY

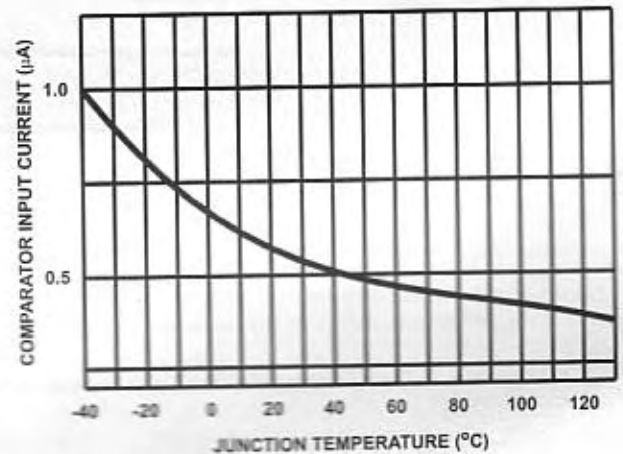


FIGURE 9. COMPARATOR INPUT CURRENT I_L vs TEMPERATURE AT $V_{CM} = 5V$

HIP4080

Typical Performance Curves $V_{DD} = V_{CC} = V_{AHB} = V_{BHB} = 12V$, $V_{SS} = V_{ALS} = V_{BLS} = V_{AHS} = V_{BHS} = 0V$, $R_{HDEL} = R_{LDEL} = 100K$ and $T_A = +25^\circ C$, Unless Otherwise Specified (Continued)

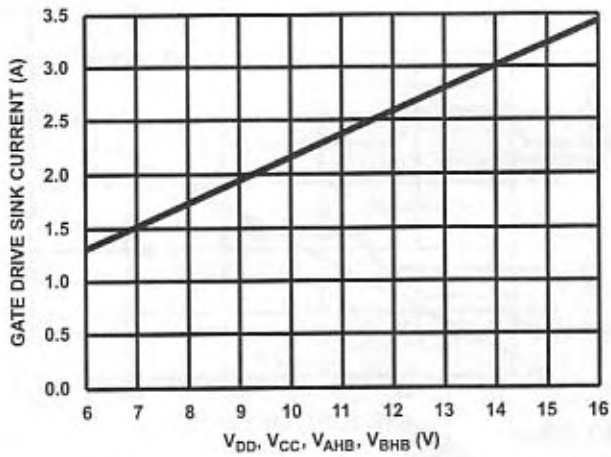


FIGURE 28. PEAK PULLUP CURRENT I_{Q+} vs SUPPLY VOLTAGE

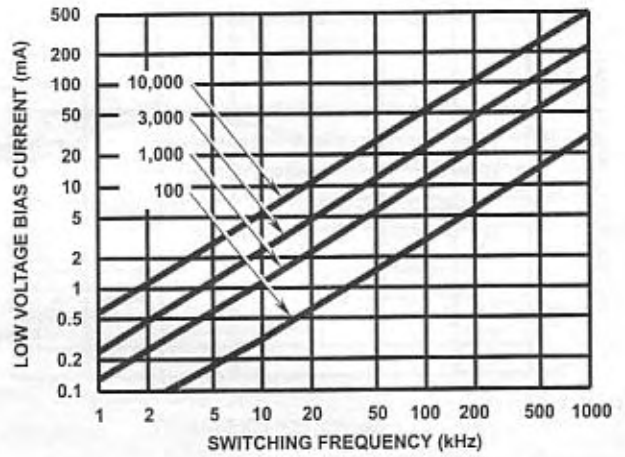


FIGURE 29. LOW VOLTAGE BIAS CURRENT I_{DD} AND I_{CC} (LESS QUIESCENT COMPONENT) vs FREQUENCY AND GATE LOAD CAPACITANCE

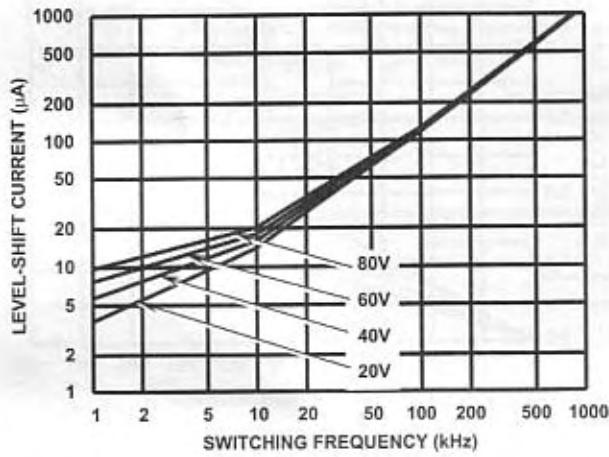


FIGURE 30. HIGH VOLTAGE LEVEL-SHIFT CURRENT vs FREQUENCY AND BUS VOLTAGE

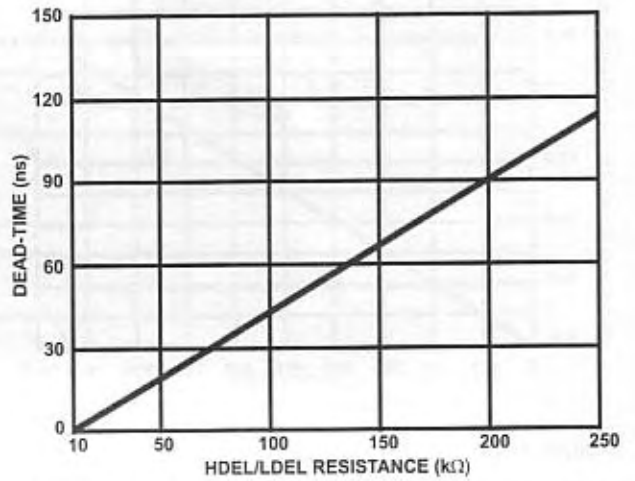


FIGURE 31. MINIMUM DEAD-TIME vs DEL RESISTANCE

HIP4080

HIP4080 Power-up Application Information

The HIP4080 H-Bridge Driver IC requires external circuitry to assure reliable start-up conditions of the upper drivers. If not addressed in the application, the H-Bridge power MOSFETs may be exposed to shoot-through current, possibly leading to MOSFET failure. Following the instructions below will result in reliable start-up.

The HIP4080 does not have an input protocol like the HIP4081 that keeps both lower power MOSFETs off other than through the DIS pin. IN+ and IN- are inputs to a comparator that control the bridge in such a way that only one of the lower power devices is on at a time, assuming DIS is low.

However, keeping both lower MOSFETs off can be accomplished by controlling the lower turn-on delay pin, LDEL, while the chip is enabled, as shown in Figure 32. Pulling LDEL to V_{DD} will indefinitely delay the lower turn-on delays through the input comparator and will keep the lower MOSFETs off. With the lower MOSFETs off and the chip enabled, i.e. DIS = low, IN+ or IN- can be switched through a full cycle, properly setting the upper driver outputs. Once this is accomplished, LDEL is released to its normal operating point. It is critical that IN+/IN- switch a full cycle while LDEL is held high, to avoid shoot-through. This start-up procedure can be initiated by the supply voltage and/or the chip enable command by the circuit in Figure 32.

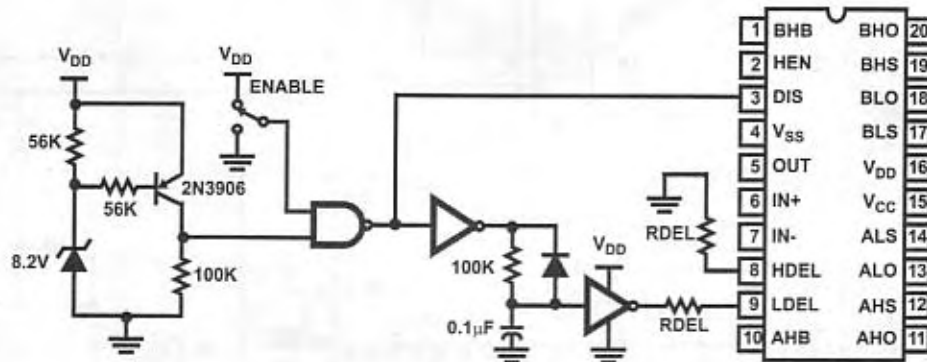
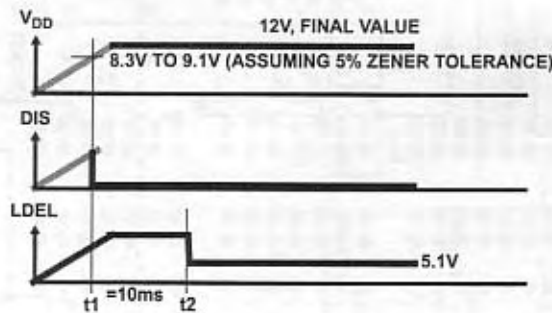


FIGURE 32.

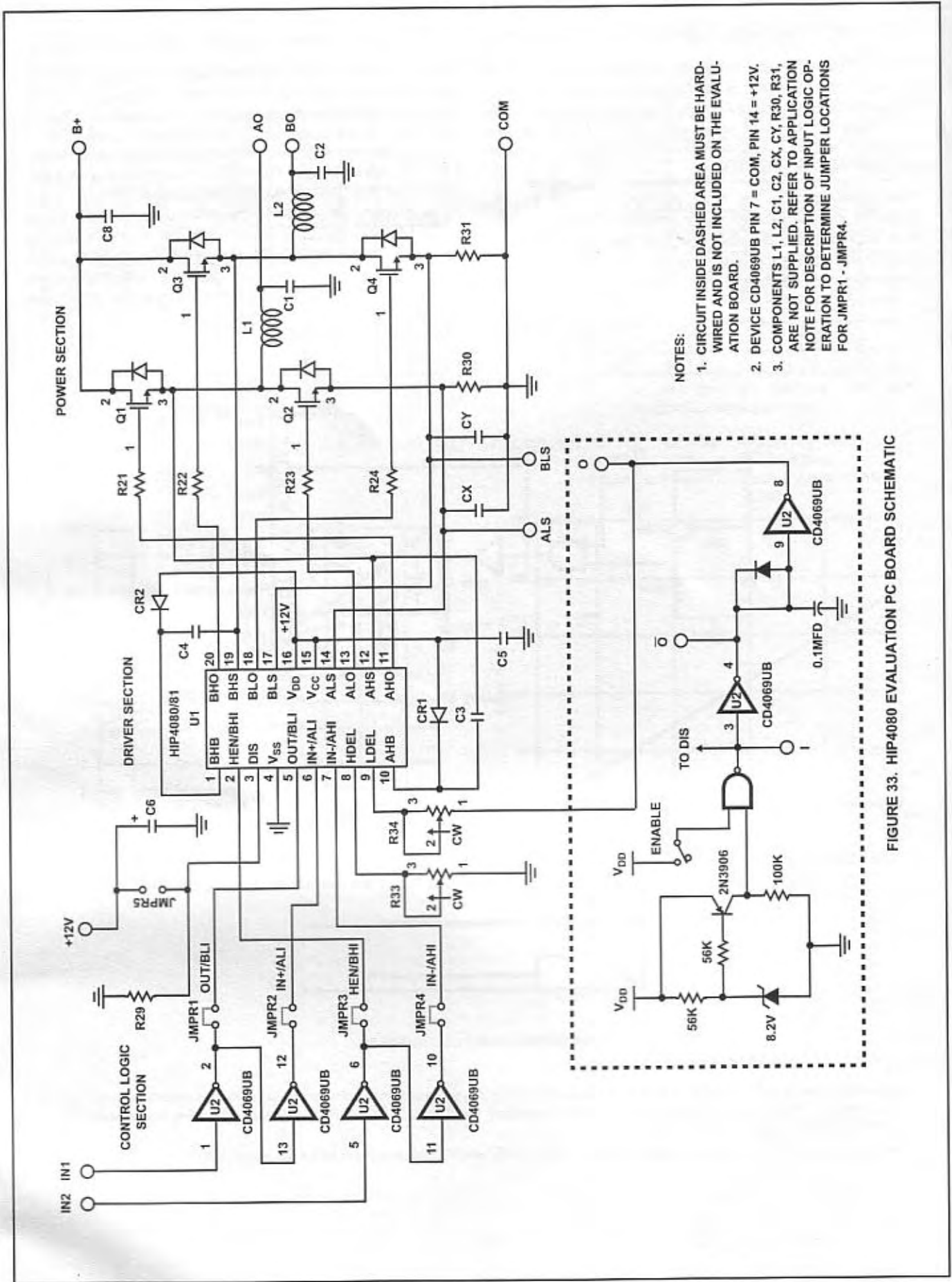


TIMING DIAGRAM FOR FIGURE 32

NOTE:

1. Between t1 and t2 the IN+ and IN- inputs must cause the OUT pin to go through one complete cycle (transition order is not important). If the ENABLE pin is low after the under-voltage circuit is satisfied, the ENABLE pin will initiate the 10ms time delay during which the IN+ and IN- pins must cycle at least once.
2. Another product, HIP4080A, incorporates undervoltage circuitry which eliminates the need for the above power up circuitry.

HIP4080



- NOTES:
1. CIRCUIT INSIDE DASHED AREA MUST BE HARD-WIRED AND IS NOT INCLUDED ON THE EVALUATION BOARD.
 2. DEVICE CD4069UB PIN 7 = COM, PIN 14 = +12V.
 3. COMPONENTS L1, L2, C1, C2, CX, CY, R30, R31, ARE NOT SUPPLIED. REFER TO APPLICATION NOTE FOR DESCRIPTION OF INPUT LOGIC OPERATION TO DETERMINE JUMPER LOCATIONS FOR JMPR1 - JMPR4.

FIGURE 33. HIP4080 EVALUATION PC BOARD SCHEMATIC

HIP4080

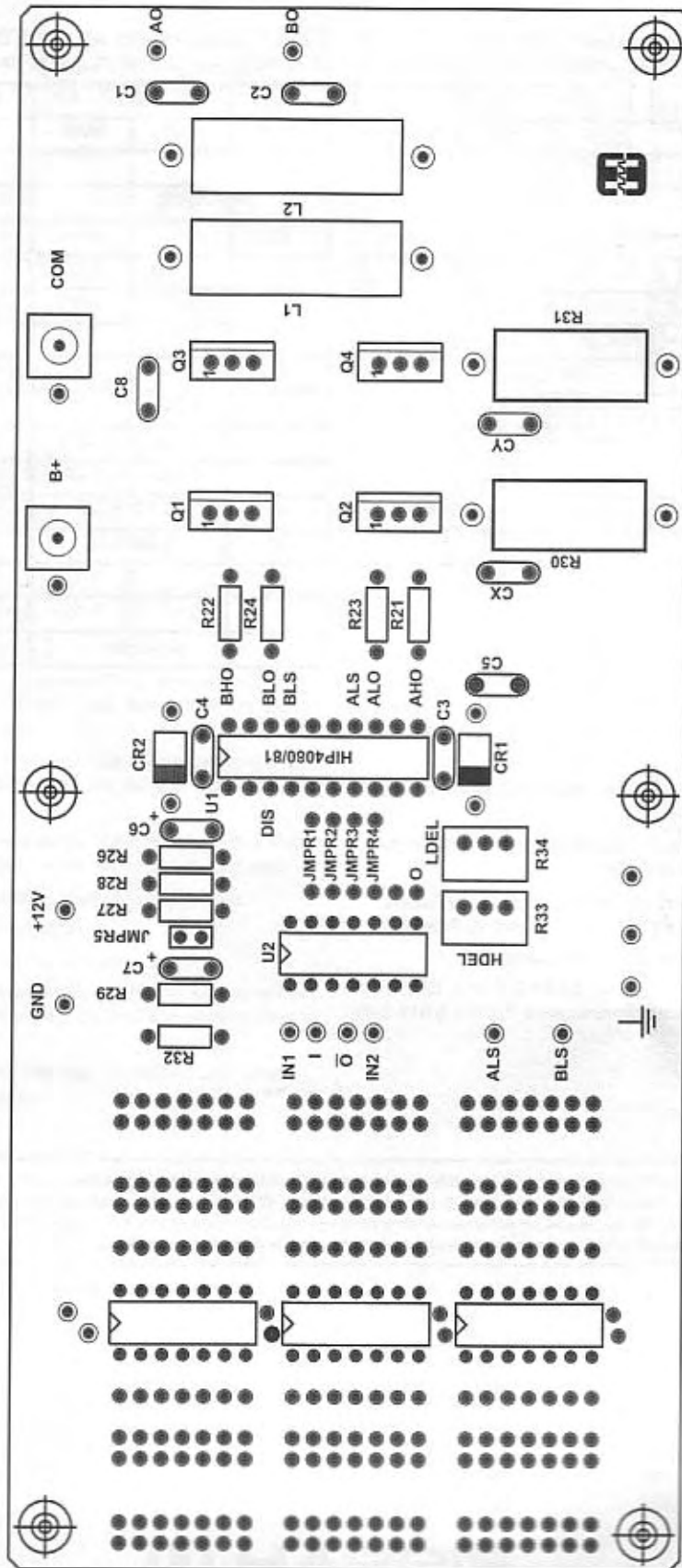
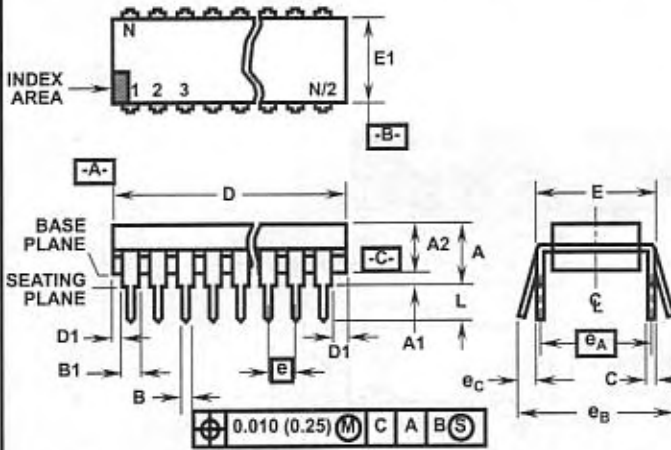


FIGURE 34. HIP4080 EVALUATION BOARD SILKSCREEN

HIP4080

Dual-In-Line Plastic Packages (PDIP)



E20.3 (JEDEC MS-001-AD ISSUE D)
20 LEAD DUAL-IN-LINE PLASTIC PACKAGE

SYMBOL	INCHES		MILLIMETERS		NOTES
	MIN	MAX	MIN	MAX	
A	-	0.210	-	5.33	4
A1	0.015	-	0.39	-	4
A2	0.115	0.195	2.93	4.95	-
B	0.014	0.022	0.356	0.558	-
B1	0.045	0.070	1.55	1.77	8
C	0.008	0.014	0.204	0.355	-
D	0.980	1.060	24.89	26.9	5
D1	0.005	-	0.13	-	5
E	0.300	0.325	7.62	8.25	6
E1	0.240	0.280	6.10	7.11	5
e	0.100 BSC		2.54 BSC		-
e _A	0.300 BSC		7.62 BSC		6
e _B	-	0.430	-	10.92	7
L	0.115	0.150	2.93	3.81	4
N	20		20		9

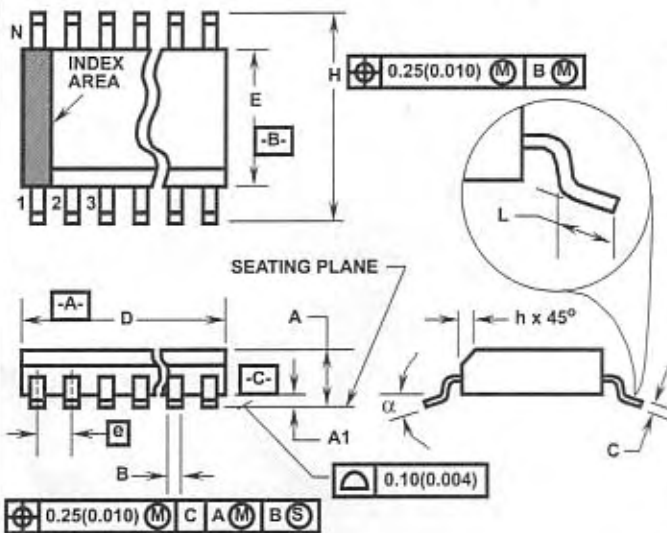
NOTES:

- Controlling Dimensions: INCH. In case of conflict between English and Metric dimensions, the inch dimensions control.
- Dimensioning and tolerancing per ANSI Y14.5M-1982.
- Symbols are defined in the "MO Series Symbol List" in Section 2.2 of Publication No. 95.
- Dimensions A, A1 and L are measured with the package seated in JEDEC seating plane gauge GS-3.
- D, D1, and E1 dimensions do not include mold flash or protrusions. Mold flash or protrusions shall not exceed 0.010 inch (0.25mm).
- E and e_A are measured with the leads constrained to be perpendicular to datum -C-.
- e_B and e_C are measured at the lead tips with the leads unconstrained. e_C must be zero or greater.
- B1 maximum dimensions do not include dambar protrusions. Dambar protrusions shall not exceed 0.010 inch (0.25mm).
- N is the maximum number of terminal positions.
- Corner leads (1, N, N/2 and N/2 + 1) for E8.3, E16.3, E18.3, E28.3, E42.6 will have a B1 dimension of 0.030 - 0.045 inch (0.76 - 1.14mm).

Rev. 0 12/93

HIP4080

Small Outline Plastic Packages (SOIC)



M20.3 (JEDEC MS-013-AC ISSUE C)
20 LEAD WIDE BODY SMALL OUTLINE PLASTIC PACKAGE

SYMBOL	INCHES		MILLIMETERS		NOTES
	MIN	MAX	MIN	MAX	
A	0.0926	0.1043	2.35	2.65	-
A1	0.0040	0.0118	0.10	0.30	-
B	0.013	0.0200	0.33	0.51	9
C	0.0091	0.0125	0.23	0.32	-
D	0.4961	0.5118	12.60	13.00	3
E	0.2914	0.2992	7.40	7.60	4
e	0.050 BSC		1.27 BSC		-
H	0.394	0.419	10.00	10.65	-
h	0.010	0.029	0.25	0.75	5
L	0.016	0.050	0.40	1.27	6
N	20		20		7
α	0°	8°	0°	8°	-

Rev. 0 12/93

NOTES:

1. Symbols are defined in the "MO Series Symbol List" in Section 2.2 of Publication Number 95.
2. Dimensioning and tolerancing per ANSI Y14.5M-1982.
3. Dimension "D" does not include mold flash, protrusions or gate burrs. Mold flash, protrusion and gate burrs shall not exceed 0.15mm (0.006 inch) per side.
4. Dimension "E" does not include interlead flash or protrusions. Interlead flash and protrusions shall not exceed 0.25mm (0.010 inch) per side.
5. The chamfer on the body is optional. If it is not present, a visual index feature must be located within the crosshatched area.
6. "L" is the length of terminal for soldering to a substrate.
7. "N" is the number of terminal positions.
8. Terminal numbers are shown for reference only.
9. The lead width "B", as measured 0.36mm (0.014 inch) or greater above the seating plane, shall not exceed a maximum value of 0.61mm (0.024 inch)
10. Controlling dimension: MILLIMETER. Converted inch dimensions are not necessarily exact.

Harris Semiconductor products are sold by description only. Harris Semiconductor reserves the right to make changes in circuit design and/or specifications at any time without notice. Accordingly, the reader is cautioned to verify that data sheets are current before placing orders. Information furnished by Harris is believed to be accurate and reliable. However, no responsibility is assumed by Harris or its subsidiaries for its use; nor for any infringements of patents or other rights of third parties which may result from its use. No license is granted by implication or otherwise under any patent or patent rights of Harris or its subsidiaries.

Sales Office Headquarters

For general information regarding Harris Semiconductor and its products, call 1-800-4-HARRIS

UNITED STATES
Harris Semiconductor
2401 Palm Bay Road N.E.
Palm Bay, Florida 32905
TEL: (407) 724-7000

EUROPE
Harris Semiconductor
Mercure Center
100, Rue de la Fusee
1130 Brussels, Belgium
TEL: (32) 2-724-2111

SOUTH ASIA
Harris Semiconductor H.K. Ltd.
13/F Fourseas Building
208-212 Nathan Road
Tsimshatsui, Kowloon
Hong Kong
TEL: (852) 723-6339

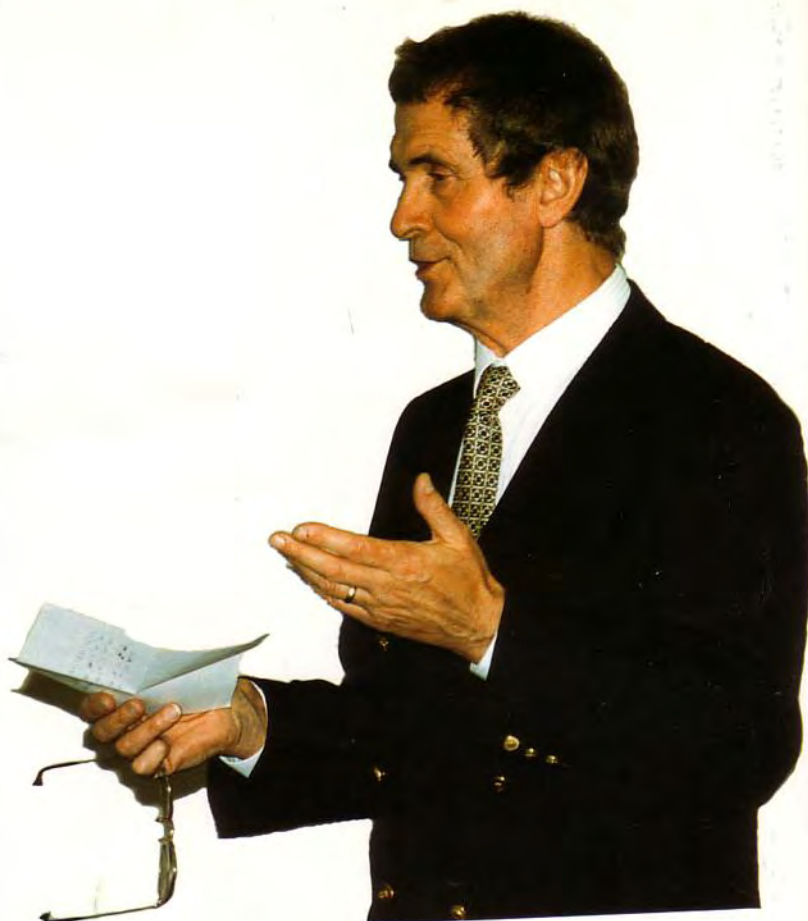
NORTH ASIA
Harris K.K.
Kojimachi-Nakata Bldg. 4F
5-3-5 Kojimachi
Chiyoda-ku, Tokyo 102 Japan
TEL: (81) 3-3265-7571
TEL: (81) 3-3265-7572 (Sales)



The LF Experimenter's Handbook has been written to meet the needs of amateurs and experimenters who have an interest in low power radio techniques at frequencies below 200kHz. Most of the techniques described are targeted at those using the 136kHz band, but they are also of great interest to readers in New Zealand and Australia with their 183kHz band and the Lowfers in the USA on 180kHz.

The bulk of the material in this book comprises contributions from experimenters world wide and covers antennas, propagation, receivers, transmitters, special modes and test equipment, and also discusses earthing techniques. Information on portable expeditions and the first transatlantic contacts are included.

All those interested in exploring the new frontier of the LF bands will welcome this invaluable and unique reference, whether they intend transmitting or are content just to receive. Anyone who bought the LF Experimenter's Source Book, which this book replaces, will want to update themselves on the very latest techniques.



the **LOW FREQUENCY EXPERIMENTER'S** handbook

£16.14P

ISBN 1-872307-65-8



9 781872 309651