Constructing the 3496Hz "D-Q" Beacon Receiver

***Brian Pease** provides constructional details for his "double quadrature" receiver which is used in conjunction with his radio-location transmitter.*

Introduction

This article gives theoretical and constructional details for a simple high performance 3496Hz long range cave radio beacon receiver which uses what I call a "Double Quadrature" detector. It is used to receive either a steady (non-pulsed) beacon signal for determining location and depth or CW (Morse Code) for passing information to the surface. The complementary beacon transmitter has previously been described in this *Journal* (Pease, 1996a).

A conventional loop antenna is used to locate "ground-zero". Searching can be aided by signal strength readings. Once ground-zero is located, depth may be measured by the traditional "null-angle" method and/or by reading the signal strength on a digital voltmeter followed by a simple calibration on the surface after the trip. Alternately, the "ratiometric" method (Gibson, 1995, 1996) involving the ratio of two strength readings taken at different heights above the surface. These last methods have allowed a complete "search – locate ground-zero – find depth" sequence to be completed in five minutes by one person on the surface for depths of 50 feet in open terrain. This can result in a happy in-cave crew if a voice down-link is used.

Receiver sensitivity is limited only by the thermal noise of the loop, which will be overcome by atmospheric or power line noise much of the time. If needed, the narrow 1Hz filter has 30dB of attenuation only 17Hz either side of the 3496Hz carrier frequency, which suppresses 60Hz power line interference. 3496Hz is not a good frequency for the UK due to a very close harmonic of 50Hz. The operating frequency can be easily changed if desired. Direct measurement of the H-field strength is also possible.

Knowledge of the conductivity of the rock can be used to improve the accuracy of the depth measurement for depths over 30-40 metres (Pease, 1991; Drummond, 1989). By placing both the beacon and receiver on the surface, a simple "depth-of-null" measurement allows easy rough estimation of average ground conductivity for any (approximate) depth.

Specifications

Theoretical ultimate range of this receiver (using headphones for searching) with my beacon with its small 2-foot diameter loop is 885 metres in "free space" assuming coaxial loops, 1Hz bandwidth, 10dB s/n ratio, and no atmospheric or power line noise. The "real world" range may be quite a bit less, however the PLL and DC meter will work *below* the noise in the 1Hz bandwidth.

Measured Specifications (12V DC)

- Sensitivity (equivalent noise H field) is 1.3nA/metre in a 1Hz bandwidth.
- Sensitivity of the rf amplifier (noise at input) is 20nV in a 1Hz bandwidth.
- Bandwidths are 1Hz and 32Hz at the -3dB points.
- Selectivity is 12Hz at -20dB points in 1Hz mode.
- Loop parameters:
  - Q=29 (25 with electrostatic shield)
  - Resonant impedance = 125kΩ (215kΩ with 2nd LC circuit)
  - Thermal noise = 54nV/1Hz bandwidth with 2nd LC circuit
  - E-field effective height = 0.12 metres
- Phase-locked loop bandwidth (-3dB) is 0.16Hz.
- PLL capture (lock) range is 0.14Hz.
- Threshold for phase-lock is about 3mV on the DC DVM.
- Threshold for the lock indicator/alarm is about 60mV on the DC DVM
- Maximum rf linear rf amplifier output is 2.75V rms.
- Maximum DC meter reading is approximately 1V DC without rf overload.
- Maximum AC meter reading is approximately 2V rms without rf overload.
- DC meter bandwidth (-3dB) is 0.15Hz with 10μF.
- Settling times: The receiver takes about 2 minutes to stabilise at turn-on. The DC meter takes 10 seconds to settle fully with 10μF.
- Power: Draws about 35mA from two 9V alkaline batteries in series, which will give several hours of life.

Circuit Description

The Loop Antenna

A previous article (Pease, 1996b) gives a more detailed circuit description than presented here. That article also included the block diagram. The loop antenna consists of one pound (450g) of #29 awg (0.286mm[1]) enamelled wire wound 18.25" (464mm) in diameter (I got 430 turns), wrapped with electrical tape and mounted on a board. I covered the winding with a (probably unnecessary) electrostatic shield.

The second tuned circuit (L1 and C35, overleaf) reduces rf amplifier overload from nearby transmitters and power lines, but should not be needed in most situations. The loop is resonated with a 1000pF Arco trimmer plus polystyrene and/or silver mica capacitors. The thermal noise of this antenna determines the maximum sensitivity of the receiver.

The RF Amplifier

The rf amplifier (U0 & U1) has been upgraded to a 3-stage design with a high impedance input using FET-input op-amps and a unique wide range gain control (R22) which varies the gain of U0 and U1A together. Counting the 40 dB input attenuator the gain can be varied from -4 to +100dB. The circuit has low input noise compared to the thermal noise of the loop.

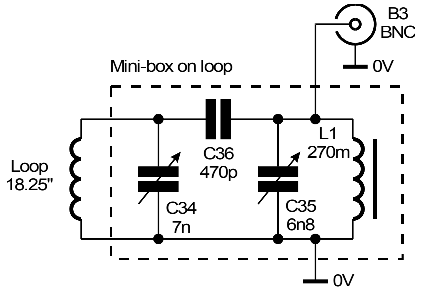
An rf overload LED (D6) utilises an Exclusive-OR (XOR) gate to indicate saturation of the rf amplifier by atmospheric noise, power line EMI, or the beacon signal. The circuit was empirically designed but the prototype works fine.

The Local Oscillator

The local oscillator (U6) uses a common 3.579545MHz NTSC colour burst crystal which is binary divided in U6 to the 3495.65Hz carrier and 437Hz audio tone frequencies. It is tuned to match the beacon frequency closely.

XOR gates U7 A & B provide the 90° phase shifts required by the detector. If a different crystal frequency is chosen then the band-pass filter (U3B) must be changed since the audio frequency will no longer be 437Hz.

[1] Nearest metric size is 0.28mm. Nearest Imperial size is 31 swg (0.295mm) or 32 swg (0.274mm).



Schematic of Optional Filter

The Double-Quadrature Detector

The narrow band frequency converting detector in this receiver is an improvement on the 8-pole commutating filter/mixer (my idea) used in Ray Cole's "Organ Cave Radio" (Cole, 1985, 1986; Stevens, 1988). It was built to solve the operational problems of my "synchronous" receiver which was used for earlier location, depth, and conductivity measurements (Pease, 1991).

This detector uses a 2-channel in-phase/quadrature direct conversion mixer (U5A & B) whose DC (base-band) outputs are low pass filtered and then up-converted (by U5C & D) to a pair of audio tones whose algebraic sum is proportional to signal strength. The great selectivity results from the fact that the 1Hz bandwidth mode rolls off at -20 dB per decade based on the 0.5Hz bandwidth of the RC filters (R1, C1, C3 and R2, C2, C4). Without the PLL the DC outputs of the two low pass filters would drift slowly with time (one is maximum when the other is zero) but in theory their rms sum will remain constant. In practice there is about 1dB variation which is inaudible but is annoying for field strength measurements. The combination of the second mixer stage, summer (U3A), and 437Hz band-pass filter (U3B) provide an audio output and allow signal strength measurement with an ordinary AC DVM. The output remains a sine wave even when the input is seriously overloaded which more or less eliminates the need for AGC, limiters, or log amplifiers while searching for ground zero.

The Audio Amplifier

The audio amplifier in the prototype was just a conventional non-inverting op-amp designed for use with my 2kΩ high efficiency headphones. For the usual low impedance (8-30Ω) stereo phones a better amplifier is needed so I designed the LM-386 circuit shown (U9). The new circuit may be more prone to feedback due to the higher currents involved.

The Phase-Locked Loop

A phase-locked loop can be easily added to the D-Q detector to solve the drifting problem by locking the receiver's local oscillator to the beacon. The PLL's main purpose is to improve the signal strength readout by allowing the use of a DC meter. It also allows for a "lock alarm" that will alert the operator when a signal is present. It has no other effect on normal receiver operation.

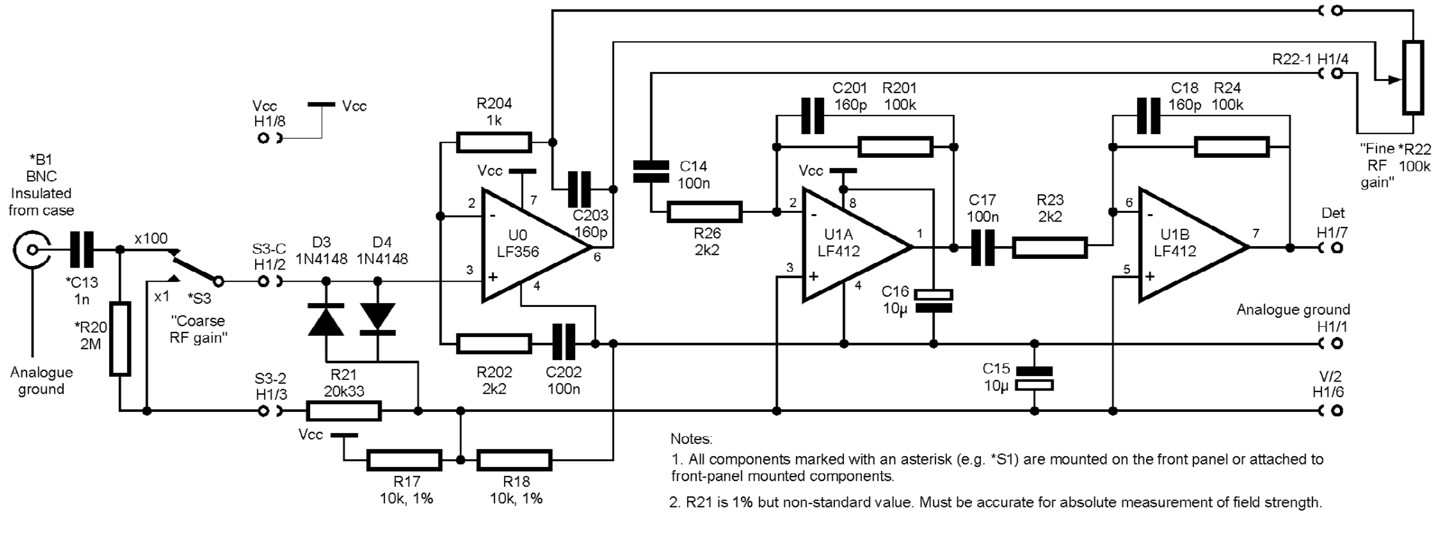
The base-band signal from one channel of the D-Q detector is connected to the input of the high gain DC coupled amplifier U4A (+60dB) whose output drives variable capacitance diode V1 which can slightly shift the frequency of the 3.57 MHz crystal. The total shift at 3496Hz is only 0.14Hz but this is more than enough. Once locked, the SIN (quadrature) signal is nulled out while the COS (in-phase) channel carries a steady DC voltage proportional to the beacon signal. Extremely narrow loop filtering allows the PLL to lock on signals that are well below the noise and interference in the receiver's 1Hz bandwidth and to give a steady readout on a DC DVM. The DC meter has two desirable features: 1) There is an inherent 3dB improvement in s/n ratio over the AC DVM (for the same receiver and meter bandwidth) since the DC meter only sees noise from one channel and 2) the DC bandwidth can be narrowed with a simple RC low-pass filter almost without limit to steady the readout. R38 & C33 give a bandwidth of 0.15Hz for an additional 8 dB improvement in s/n ratio. This steady (positive polarity) reading is the best proof that the receiver is phase-locked. In poor conditions, the DC readout is superior to the AC meter, although neither exhibits "drift" while the receiver is locked. My prototype has a built-in digital panel meter that shares the receiver's power source, but requires a differential amplifier to isolate the grounds.

Lock Indicator and Audio Alarm

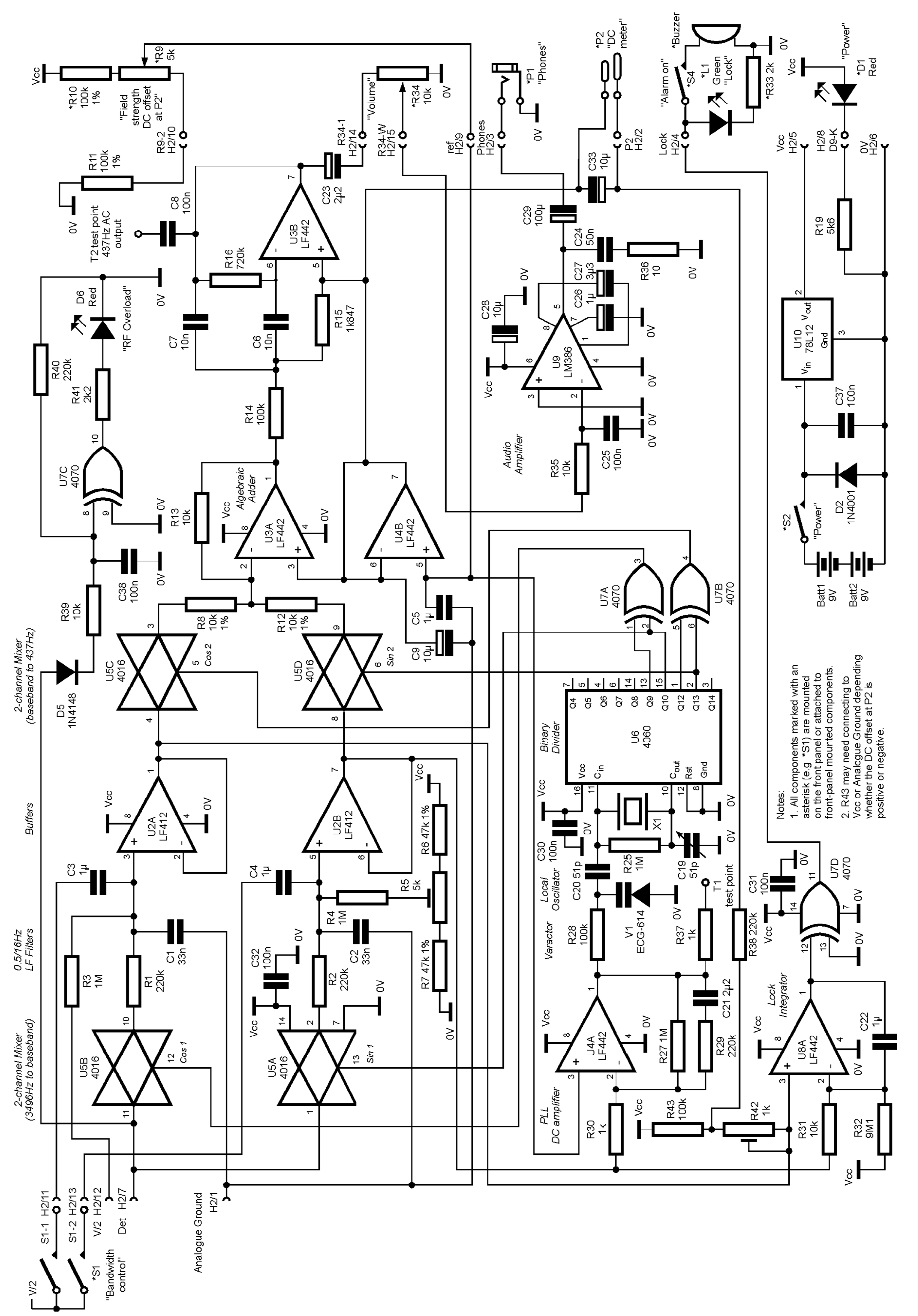
To make waiting on the surface less boring I added a circuit to indicate when the receiver phase-locks on a beacon. U8B is an op-amp integrator with a differential input that monitors the base-band (DC) output of both channels. When the in-phase channel rises a few millivolts (positive) above the quadrature channel and holds for several seconds, the integrator output will rise high enough to trip XOR gate U7D and light the "locked" LED and sound a loud alarm if desired. Its threshold is set higher than the minimum PLL lock-on signal in order to reduce false alarms. It is not foolproof but it has worked fine in field tests so far.

Construction

Ian Drummond laid out a 3-part printed circuit board containing the beacon, original receiver rf amplifier, and receiver. Four boards were made and three are actually working. We may offer some updated boards for sale if there is enough interest. I plan to construct an experimental 874Hz receiver with my board since my prototype 3496Hz unit works just fine.



Schematic of RF Amplifier Board



Schematic of Main Board

For this receiver to work properly, digital noise, especially the 3496Hz local oscillator signal, must be kept out of the analogue circuits, particularly the rf stage, otherwise adjusting the rf gain will affect the detector null. With the PLL circuit, any 3496Hz or sub-harmonic leakage may cause "lock-up" on the receiver's own local oscillator signal at high rf gain settings and possible variation in accuracy at different gain settings. The entire radio must be shielded to prevent feedback to the loop antenna. I don't know if the loop's electrostatic shield is necessary, but I do get extremely deep nulls at close range (>70dB) and have no "hand" effects. The second tuned circuit (if used) should be shielded and mounted on the loop to isolate it from the digital circuitry. I can place my 2kΩ headphones and the entire receiver, at maximum rf and audio gain, into the centre of the loop without feedback or noise of any kind.

I built the rf prototype amplifier on its own Radio Shack board and shielded it, along with the input connector (which is not grounded to the case), attenuator, and rf gain control, from the rest of the receiver. The bypass capacitors for $+V_{CC}$ and V/2 (C15 & C16) are also included. I used a ten turn potentiometer with calibrated dial for rf gain. As a precaution, all analogue grounds are brought individually to a single ground lug bolted to the partition separating the rf amplifier from the main board. Oscillation is always a potential problem with 100dB gain. I placed a shield of grounded foil between the input circuits (B1, S3, etc.) and the rf amplifier circuit board to eliminate some obvious feedback at maximum gain. I also used a very short coax to connect pin 3 of U0 to the input circuits, with the shield connected only at the input end.

The layout of the main board is not critical except to keep digital signals away from the audio amplifier. Again, in the prototype all analogue grounds are brought individually to a lug on the same bolt holding the rf amplifier ground lug. The prototype used an RS 2 × 3 × 5 inch (50×75×125 mm) aluminium mini-box for overall shielding and a belt clip for hands-free operation. Ian's custom PC boards require a larger box.

If the PLL circuits are not being installed then C20 is replaced by a 30pF capacitor connected from pin 11 to ground. V1 along with all parts associated with U4A, U7D, U8B, and the DC DVM output are not installed.

The D-Q detector circuit must be carefully adjusted to null out the 437Hz tone (when no signal is present). If the PLL is not installed then you may have to replace R8 & R12 with a pot to equalise the gain in the two channels to minimise fluctuations in the AC output level when drifting phase causes the signal to shift from one channel to the other. I put the "null" control (R9) on the front panel with a knob and the "null balance" pot (R5) on the circuit card but accessible from outside with a screwdriver. "Null balance" should only need touching up once or twice a day when temperature changes. It will pay to use 1% resistors (or matched pairs) for all three DC divider networks. The actual values are not critical. The "null balance" pot should be centred before installation to aid in the initial tune-up.

The receiver will work directly from a single 9V battery without a voltage regulator if desired, but there will be significant drift of the null as the voltage drops along with small changes in gain.

Initial Tune-up

1. If the PLL circuit is installed, break the loop by removing the 100kΩ resistor (R28) from pin 1 of U4A and connecting it to V/2.
2. Turn on the receiver while monitoring current drain from the battery. Mine is 35mA at 12V DC. Do not connect the antenna.
3. Check all three voltage divider circuits for a nominal value of ½ the supply voltage.
4. The output of each op-amp should also be about ½ of the supply voltage. If the PLL circuits are installed, the output of U4A is acceptable if it is within 1-2V of V/2 and varying.
5. You should hear a 437Hz tone in the earphones. With the rf gain switch in the "low" position and the rf gain control at minimum, alternately adjust "null" and "null balance" controls until a deep null is found, leaving only noise. If you run out of adjustment range, it may be necessary to trim one of the voltage divider resistors.
6. Now put the rf gain switch in the "low" position and the rf gain control to maximum. The output noise level should increase, especially in the 32Hz bandwidth mode. Now connect the antenna.
7. Tune up the loop tuned circuits by using your beacon signal while monitoring the AC output of the rf amplifier directly (if possible). Keep rf amplifier output below 1 volt rms to avoid saturation. The 437Hz audio tone will be steady if both channels are working. With the beacon off, in the 1Hz mode at high rf gain you should now be able to detect individual lightning strikes. Atmospheric noise is loudest at night and least in the morning. It is also loudest in the summer and least in the winter.
8. Match the receiver to the beacon frequency by first receiving a fairly strong beacon signal. If the PLL circuits are installed simply monitor test point T1 with an *analogue* DC voltmeter while adjusting C19 to lower the beat frequency as close to zero as possible. Without the PLL circuits, monitor one of the DC outputs of U2 or temporarily disconnect R8 and monitor "pulsing" audio.
9. Reconnect R28 and/or R8. A receiver with PLL should lock on the beacon signal. C19 can be touched up to "centre" the voltage at test point T1 at V/2. The lock LED L1 (not to be confused with the coil L1) should light. The sensitivity of the lock indicator is adjusted by R32. Raising its value increases weak signal sensitivity but will increase false triggering from noise at high rf gain.
10. A small "offset" will exist between the audio and DC DVM nulls with the receiver adjusted as in step 5. Adjust R42 to null the DVM. It may be necessary to move the connection of R43 from B+ to the analogue ground to make this adjustment depending on whether the offset is positive or negative.

Calibration of the rf gain controls and "absolute" calibration is beyond the scope of this article. Calibration of the controls is necessary for measuring conductivity as the relative strengths of a peak and a deep null must be recorded. Absolute calibration allows depth by field strength to be measured in "real time" and allows one calibration point to be used for widely different depths. No calibration is necessary for the "ratiometric" method of depth measurement as the two numbers will always be similar enough to be recorded without changing the gain settings.

Operation

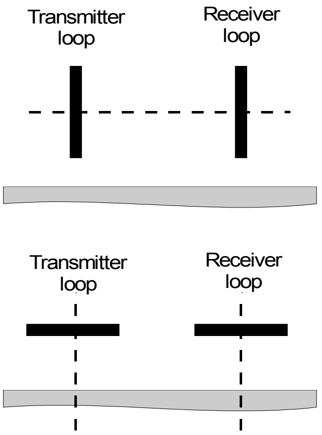
General

After several minutes' warm-up, I carefully null the receiver, using both null controls, at minimum rf gain and without the antenna. The front panel null control will need adjustment occasionally during the day using the same procedure. With the PLL receiver I just set the loop on the ground; turn on the alarm; switch to 1Hz bandwidth; and then increase the rf gain as much as possible without false alarms. I am then free until the alarm sounds when the continuous beacon signal comes on. If the signal is strong I will use the wide bandwidth mode while searching due to its faster reaction time.

Depth by Null Angle

The details of locating a ground-zero and measuring depth by the "null angle" method have been given too many times to repeat here. I use a table giving depth-to-horizontal distance ratios for each 0.5° of loop tilt from vertical (Pease, 1997b). Null angles of 25° to 40° from the vertical should give the best results. With this receiver one has the option of measuring the relative strengths of the horizontal and vertical fields separately and calculating the null angle as an arc-tangent. This seems to give more accurate results than direct measurement when the null is very shallow (<20 dB).

Depth by Absolute Field Strength



Loop Geometry
Top: Coaxial (loops on edge)
Bottom: Horizontal (loops on side)

In its simplest form, depth-by-absolute-field-strength requires calibrating the beacon and receiver on the surface after the trip. It also requires some sort of rigid frame for the beacon loop. After locating ground-zero, set the receiver's loop horizontally on the ground and switch to the 1Hz mode. Connect the DC DVM and set the rf gain controls so that the rf overload LED does not light. Record the maximum reading and the exact rf gain and switch settings. Later, set up both loops coaxially on the surface as shown in the illustration on this page. With this geometry the ground has little effect on the signals (at moderate spacing over limestone anyway) so the result will be close to the free-space value. Use the same receiver settings recorded earlier. Now simply adjust the spacing to duplicate the reading obtained at ground zero and measure the distance to obtain the depth.

Coplanar surface calibration, with both loops lying on the surface, is possible, but is restricted to short spacing (perhaps 30-50 metres) as the ground has more effect on the signal than in the coaxial arrangement. Remember that the received signal strength is exactly one half that obtained with the coaxial arrangement.

Depth by Ratiometric Methods

Ratiometric depth measurement is perhaps the simplest method overall with no calibration or angle measurements required (Gibson, 1995, 1996). Once at ground-zero (precise location is not essential) one simply records the field strength, V_1, with the receive loop horizontal then raises the loop a known height, H, (5% of expected depth is a good minimum) and records the strength again, V_2. Since it is not necessary to adjust the receiver's gain between readings, and only the ratio of the numbers is used, no calibration is required. The rf amplifier must not be overloaded and the beacon signal must remain constant for accurate results. The calculation is a variation of the free-space cubic fall-off equation:

$$D = \frac{H}{\left(\sqrt[3]{\frac{V_1}{V_2}}\right) - 1}$$

The conductivity of the rock will cause errors in all three of the depth measurement techniques, but good results should be obtained up to a depth of 30-60 metres with any of the methods. In homogeneous (uniform) earth, the null-angle and ratiometric methods should always give a value *less* than the actual depth, while the field-strength method should always give a value *greater* than the actual depth. As depth increases, the spread between the absolute and ratiometric methods will increase, but the actual depth should always lie between them! At great depths the average of the two values will be closer to the actual depth than either value alone. I have successfully simulated these effects by using a computer program that calculates the effect of conductivity on the strength and direction of the beacon's magnetic field (Pease, 1997a). This article describes all three depth methods and quantifies the effect of conductivity. Using two or three methods is also a good way to pick up careless errors, even at shallow depths!

Further Information

Feel free to contact me for any details of construction, calibration, or operation. I also have an idea for improving skirt selectivity which I will try when I build my 874Hz unit. I am also available to do locating work using this gear. My postal address, phone number and e-mail address appear at the end of this article.

Parts List

Resistors

(¼W, 5% carbon film except as noted)

R1, R2, R29, R38, R40	220kΩ
R3, R4 , R25, R27	1MΩ
R5	5kΩ trim pot
R6, R7	47kΩ 1% metal film
R8, R12, R17, R18	10kΩ 1% metal film
R9	5kΩ linear multi-turn pot
R10, R11	100kΩ 1% metal film
R13, R31, R35, R39	10kΩ
R14, R24, R28, R43, R201	100kΩ
R15	1847Ω
R16	720kΩ
R19	5.6kΩ
R20	2MΩ 1% metal film
R21	20.33kΩ (1% metal film trimmed to exact value)
R22	100kΩ (10-turn linear pot with a calibrated dial)
R23, R26, R41, R202	2.2kΩ
R30, R37, R42, R204	1kΩ
R32	9.1MΩ
R33	2kΩ
R34	10kΩ audio taper (i.e. logarithmic) pot
R36	10Ω

Capacitors & Inductors

(All capacitors monolithic ceramic 0.1" spacing, except as noted)

C1, C2	33nF
C3, C4, C5, C22	1μF
C6, C7	10nF
C8, C14, C17, C25, C30, C31, C32, C37, C38, C202	100nF
C9, C15, C16, C28, C33	10μF Tantalum 16V
C13	1nF
C18, C201, C203	160pF
C19	20 to 60pF trimmer
C20	51pF
C21	2μF ceramic (could be two 1μF in parallel)
C23	2.2μF Tantalum 16V
C24	50nF
C26	1μF Tantalum 16V
C27	3.3μF Tantalum 16V
C29	100μF electrolytic
C34	7nF
C35	6.8nF
C36	470pF
L1	270mH shielded (Mouser)

Note:

1. There is no C11 or C12.
2. C34 and C35 are approximate values. They must be varied to resonate.

Continued on page 17

Low-noise RF Amplifier uses FET Cascode Input

***John Hey** presents an alternative RF amplifier for the G3TDZ cave radio. Much improved results are reported.*

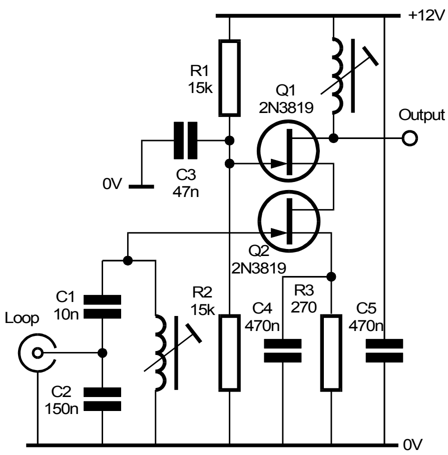
Introduction

In the original design for the G3TDZ cave radio (Hey, 1995a, b), the ultra-low noise industry standard op-amp, the OP-37, was chosen for the receiver RF amplifier. Op-amps are very easy to design with but, by their very nature (a box full of transistors with all their junctions), must generate a fair degree of noise. What an instrument engineer considers "low noise" might be very different from a radio engineer's point of view. Because it was thought that the background noise was a little high, even without an antenna connected, an alternative was tried. That alternative is described in this article.

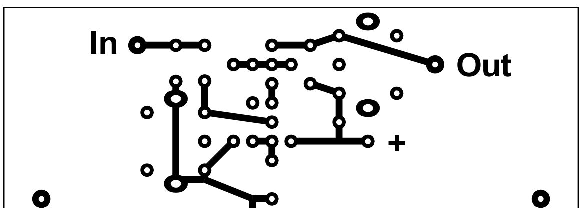
The New Circuit

The well known old favourite, the Cascode amplifier, was constructed using a pair of junk-box FETs. The Cascode has modest gain and is very stable. The first surprise was that the gain is exactly the same as that of the op-amp, that is 48dB. Before fitting the new circuit, the noise level was measured at the top of the volume control. It was 15mV. With the new amplifier fitted, the noise had dropped to under 1.2mV; this is a reduction of 22dB.

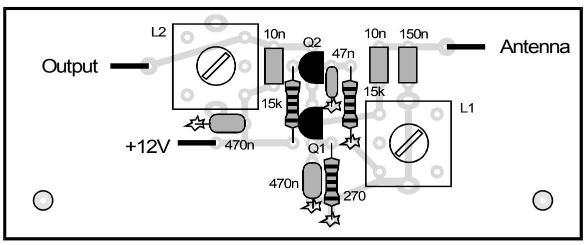
This experiment had been carried out on my surface radio. Would the underground set prove as exciting? A second amplifier was built and the same tests carried out. With the original circuit fitted, the noise level was 9mV – better than its mate, but still far behind the 1.2mV result. Before the substitution, the signal to noise ratio was measured to show 1:1 S/N at 2μV. The new RF amplifier was fitted and the noise level dropped to 0.75mV, an improvement of 21.6dB. The S/N at 1:1 was now 0.35dB.



Schematic of New Pre-amp



PCB Layout – actual size
(Upper surface is solid ground plane countersunk around holes except where component leads connect to ground.)



Component Overlay

Results

In the quiet of one's radio shack, a 22dB improvement in noise figure is a real bonus; in a noisy cave situation, it enables the audio to be turned up further. Receiver sensitivity remains much the same as before and radio noise is likely to exceed internal noise so the full 22dB improvement will not be realised in practice. It is, however, always in the best interest to make improvements wherever possible and the new pre-amp does just this.

The new pre-amp was first used underground during the March field meeting but not at any significant depth. It is hoped to conduct some tests at a significantly greater depth during the Summer.

Parts List

Resistors (all ¼W, 2%)	
R1, R2	15k
R3	270R
Capacitors	
C1	10n
C2	150n
C3	47n
C4, C5	470n
Inductors	
L1	
L2	
Semiconductors	
Q1, Q2	2N3819

References

Hey, John (1995a) *The G3TDZ Cave Radio*, CREGJ **22**, pp12-16.

Hey, John (1995b) *The G3TDZ Loop Antenna*, CREGJ **23**, pp21, 24.

John Hey & Nigel Lovell Testing the G3TDZ Cave Radio in Jug Holes

CREG

White LEDs: Introducing a New Generation

After years of waiting, the blue LED is finally with us. But closely associated is something of far more potential interest to the caver. ***Steve Clark*** *reports on the first of the white LEDs.*

A new type of LED has just come onto the market, and should prove to be a very interesting light source. Marketed as a white LED, it is, in fact, a high intensity blue device "driving" a phosphor. The phosphor appears to be layered directly onto the mounted die and then encapsulated in clear epoxy. Most of the blue light is used to energise the phosphor which then re-radiates the energy at a range of lower frequencies – a broad spread peaking at around 550nm (compared with 400nm for the blue).

I have obtained samples of the 3mm version (this is the only size currently available – it'll be a few months before the more robust 5mm package and surface mount chips will be available) and they look promising. The specification is given below, with details of price and supplier.

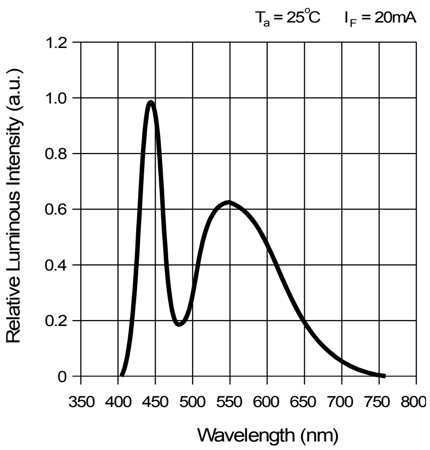
These devices are neither the brightest nor the most focused LEDs on the market. However, what makes them useful is that they allow the eye to perceive colours reasonably accurately. A single device running at 20-25mA (68Ω resistor and 4 × 700mA-Hour AA NiCds – over 24 hours light per charge!) is more than adequate for seeing around the house – no external reflector is needed, and there's no filament to fail. The beam is wide (60° cone) and appears white with a blue tint. I have converted a "cheap" headlamp (Elastic headband, 3 × AA alkaline cell holder, and small slide switch on battery pack) which gives a well focused central spot – capable of picking out cobwebs in the corner of a ceiling at five metres! Caving gave me up a few years back. so I haven't tested this lamp in "true" darkness yet, but I know a man who can! A suggestion for an emergency light includes a pair of LEDs in series with 82Ω. This will run for over 20 hours from an alkaline PP3 battery. The Polaroid "Polapulse" batteries produce 6V and, with a 100Ω resistor, could provide an almost two dimensional, compact light. The shelf life of reputable brands of alkaline batteries is quoted as five years – check packaging before purchase!

The devices are manufactured in Japan by Nichia – the only company that currently has the technology for true high brightness blue, green and blue-green (lovely colour) LEDs. All such devices on the market (including the £8+ devices that RS lists) use the Nichia die. PlusOpto can also supply these devices and are much more competitive than RS. Hewlett Packard hope to have a competing technology available later this year which may help prices to drop. One question which I do not yet have the answer to concerns the phosphor's half-life. If it's similar to fluorescent lamps, it would manage around 10,000 hours to 80% brightness – still a good few years caving!

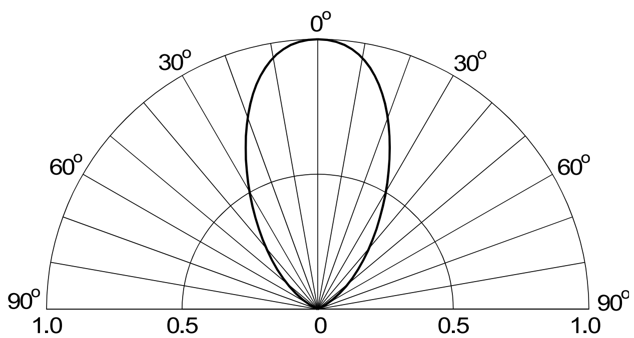
I hope to include more details of these and other high brightness devices, with comparative details and application notes in the next *Journal*.

Steve's lamp was tested at the Field Meeting and attracted very favourable comments – Ed.

CREG



Luminous Intensity



Polar Plot

Specification

White (Blue) LED - Nichia NSPW 310AS

Half angle	30° to half-power points
Max. DC current	25mA
Max Pulsed current	80mA
Operating temperature	-20°C to +80°C
DC Forward voltage	3.6V nominal, 4.0V max at 20mA.
Luminous Intensity	400mcd typical at 20mA.
Efficiency	5lm/W
Colour temperature	8000K, Chromaticity co-ordinates x0.29, y0.30, RA 85%. (More about these in the next *Journal.*)
Supplier	plusOpto Ltd.
Address	Leigh Wharf, Canal Street, Leigh, Lancashire.
Telephone	01942 - 671122, Fax 671133. Contact Andrew Heaps.
Price	1 to 10 Devices may be purchased at £3.60 each plus 39p post plus VAT.

Ground Shorts with a Single-wire Telephone

*What happens when the conductor of a single-wire telephone is accidentally shorted to earth? **David Gibson** analyses where the current goes, and shows that shorts are not necessarily a problem.*

Introduction

There has been some discussion recently (e.g. Gibson, 1997a) about the relative merits of low-impedance and high-impedance single-wire telephones. The discussion centred around two problems, namely i) the effect of a poor earth and ii) the effect of a short to ground, both of which cause signal attenuation. The debate concerned whether and how the operator would be aware of such attenuation, and what he could do about it (see box below).

To model the behaviour of a SWT in the presence of a poor earth or a line short, it is necessary to obtain expressions for the earth resistance and to understand where the current flows when a "short" occurs. This article is an attempt to provide a simple mathematical model.

A Typical Configuration

Figure 1 shows a typical SWT set-up. R_b represents the 'human body' contact resistance between the casing of the telephone and the earth or rock – the resistance between wet palms pressed onto a conducting surface can be less than 10kΩ. If the telephone is connected to a metal stake then this "contact" resistance is very low and can be ignored. R_b is present at both the transmitter and receiver. In addition, the input resistance of the receiver is represented by R_i. The resistance of the wire itself, and the output resistance of the transmitter, are negligible. (The resistance of 7/0.2mm equipment wire is around 550Ω/km. The only time this is significant is if there is a long section of wire in contact with the ground – see "attenuation along a bare wire" in Gibson, 1997b)

Suppose the wire touches the ground at some point along its path. Clearly, even with a "short" at point P, some current will still flow directly from the transmitter to the receiver, and we can represent the current flow in the earth by the resistances R_{e1}, R_{e2} and R_{e3}. To make the analysis easier, it is perfectly valid to perform a 'star-delta' transformation, and to represent the three resistances in a 'star' as shown in figure 2.

It will be clearer to represent figure 2 in a schematic form. Performing a topological transformation results in figure 3 where point P shorts to the electrical 'ground' through resistance R_P, which is related to the size of electrode P. Clearly if electrode P has a high enough resistance compared with electrodes T and R then the signal will not be attenuated to any great degree. We need, therefore, to find a way of expressing the resistances of the various current paths.

Calculating the Electrode Resistance

When two earthed electrodes are separated by a distance much greater than their size, we know that the resistance between them depends only on the electrode sizes, and *not* on the distance itself. (Gibson 1994a). The resistance between two hemi-spherical electrodes of radius a buried on the surface of an infinite half-plane is

$$\frac{1}{\pi\sigma a} \tag{1}$$

and the resistance between two rods (strictly speaking, two prolate hemi-ellipsoids) is

$$\frac{\ln(4l/d)}{\pi\sigma l} \tag{2}$$

(Gibson, 1994b) where l and d are the length and diameter of the rod ($l >> d$) and σ is the conductivity of the rock.

These formulas apply to the resistance between two identical electrodes. When the electrodes are of different sizes, it will be useful to be able to represent the resistance of a single electrode, such that the overall resistance of the path is found by summing the resistances. The resistance of a single electrode does not, of course, have any physical meaning. (Geophysicists sometimes define it by taking the second electrode to be

Single-Wire Telephones

Earth-Return Telephones

The traditional single-wire or earth-return telephone has been around since the earliest days of telegraphy in the last century. A cavers' telephone, using an earth-return path through the metal case of the unit and the human body, was first described by Neville Michie (1974) and reported by Were (1984). In 1988 Stuart France described an adaptation which is in regular and indispensable use by South Wales CRO.

More recently Nigel Lovell (1993) has described a high-impedance version with some advantages over the basic Michie design. Lovell's units have been successfully used on expeditions. (Monaghan, 1993).

The Michie-Phone

Neville Michie's design, and subsequent adaptations, all operate on the principle of a simple voltage follower. The Tx boosts the mic. signal and presents it to the line. At the Rx, this voltage is presented to a unity-gain buffer and on to the speaker. The input R of the Rx forms a potential divider with the series R of the body/earth return path and so, for the unit to work satisfactorily, the body/earth R must not be too high relative to R_{in} which, in Michie's case, is ≈ 60kΩ.

Lovell's High-Z 'Phone

A disadvantage of the Michie-phone arises if the operator does not understand that he has to have a good earth. Nigel Lovell cured this problem by increasing R_{in} to 10MΩ so that a much higher body R could be tolerated. The performance of Lovell's units is impressive – there is no need for the operator to attempt to earth himself at all, even when standing in rubber boots on a dry, sandy floor. In fact, the phones work even if both operators simultaneously jump into the air. This can be attributed to the small capacitance between the body and the ground giving an a.c. path for the audio frequencies.

Pitfalls

The problem with a high-R input is that it can be shorted out by water. In practice, we have not noticed this but, even if true, the receiver cannot function any *worse* than a conventional Michie-phone. However, Michie has raised the point that a low-R system at least has the advantage that the operators will be *aware* of poor earths and can correct them. With a high-R input they might not be aware, during reception, of a poor earth and would then not be able to transmit if the resultant high-R *output* was shorted by a leakage path to earth some way along the wire. Clearly, if the operators are told that they are not being received loud and clear, they must improve the earthing.

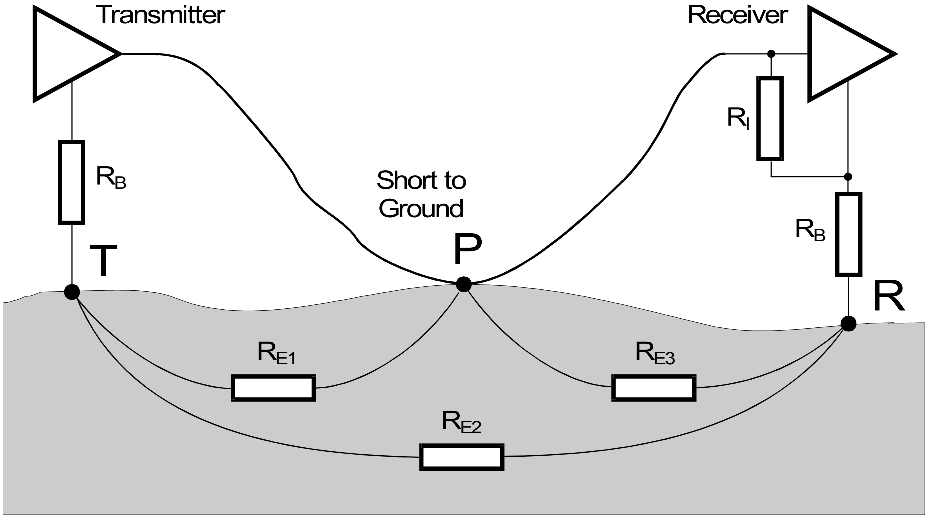


Figure 1 – A representation of a typical single-wire telephone
The human body resistances R_b are shown, together with the currents in the rock when the line is accidentally shorted to ground.

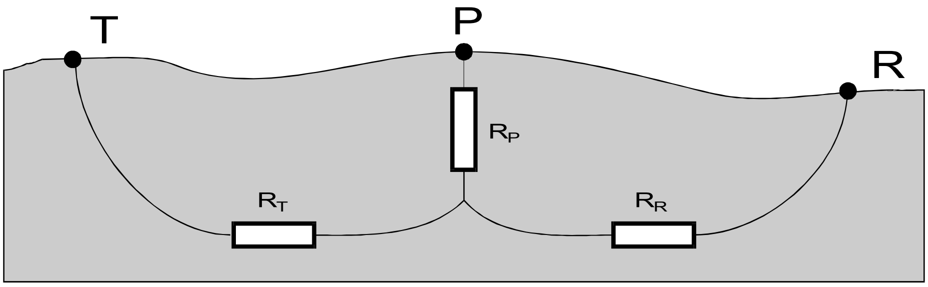


Figure 2 – An alternative representation of the earth-currents due to a short at P
The transformation from a "delta" to a "star" representation makes the analysis easier.

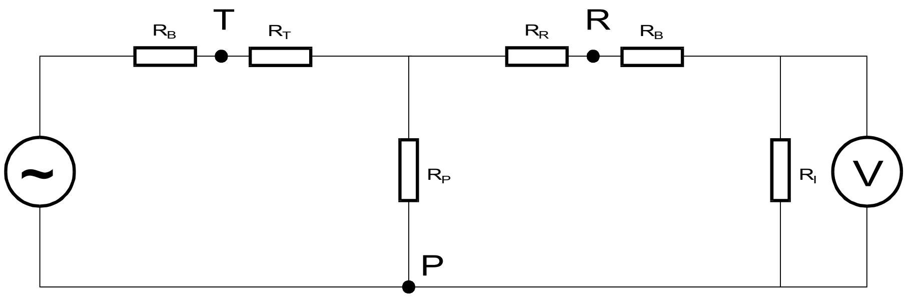


Figure 3 – A schematic version of figures 1 and 2
with human body resistance R_b, amplifier or 'voltmeter' input resistance R_i and electrode resistances R_T, R_P and R_R. This interpretation must be treated with care.

at infinity, but this is not appropriate in our situation). By an argument involving equipotentials we can define, *for our purposes*, the effective resistance of a single electrode as half the figures given in (1) and (2). The resistance between two dissimilar hemispherical electrodes thus becomes

$$\frac{1}{2\pi\sigma}\left(\frac{1}{a_1}+\frac{1}{a_2}\right) \tag{3}$$

You might think that it is now a simple matter to write the three resistances R_P, R_T, R_R in terms of the resistance of the individual electrodes since, by inspection, these would *seem* to correspond to the *R*s in our model. But "seem" is not sufficient justification, and we must proceed cautiously.

The resistance between two electrodes depends on the path taken by the current. If we apply a voltage between points *T* and *P* in figure 2 then the current flow follows a simple dipole rule, and we can work out the resistance. If, in addition, we *force* a current flow between *P* and *R* (e.g. by connecting a resistance between them) then this may, we surmise, distort the field lines, thus altering the overall current flow and the resistance of the rock. In all cases involving multiple electrodes it is advisable to calculate the resistances from first principles, just to be on the safe side.

High Resistance Receiver

As a first step in the process, we can easily work out the resistance of the current paths assuming that there is *no* current flow between *P* and *R*. This would be the case if we used a very high resistance input to the op-amp. In fact, the arrangement then becomes that of figure 4, where we are injecting a current at two points (*T* and *P*) and measuring a potential (with no current flow) between two further points (*P* and *R*). Geophysicists will recognise this as a standard tool for resistivity measurements.

Suppose *T*, *P*, *R* are hemispherical electrodes on an infinite half-plane. Appendix 1 shows that, in this case, we can write

$$\frac{V_{PR}}{V_{PT}}=\frac{R_P}{R_P+R_T} \tag{4}$$

where R_P, R_T are the resistance of the single electrodes *P* and *T* as discussed above.

This confirms, as we would expect, that the resistances in figure 3 correspond to the electrode resistance, for the case when no current flows between *P* and *R*. The expression does not include the resistance of the receiver electrode because the receiver had an infinite input resistance. Notice also, that the attenuation does not depend on the distance between any of the electrodes – only on their sizes. (provided that they are much further apart than their diameters).

Low Resistance Receiver

If the receiver does not have a high input resistance then current will flow from *P* to *R* through the wire, and from *R* into the earth. This will produce an electric field which has the possibility of distorting the current flow. The analysis of appendix 2 shows that, in fact, our concern was unjustified and that the system *can* be described by figure 3.

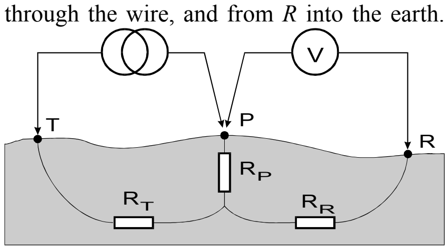


Figure 4 – A "shorted" SWT can be analysed as a conventional resistivity array

This is very useful because we can now analyse the relative performance, in the presence of a 'short', of either a high-resistance or a low-resistance receiver.

Modelling a Break in the Wire

It should be obvious, from figure 4, that we are only a short step away from the standard geophysics application of a four electrode array. Suppose the wire is broken at *P* and the two ends make separate contact

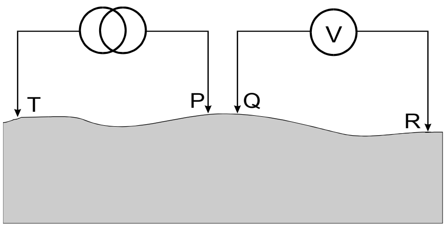


Figure 5 – A break in the wire makes the system look like a 4-terminal array

with the ground at *P*, *Q* – figure 5. If we assume that distance *PQ* is much less than the distances to points *T* and *R* and that the receiver draws no current, then we can obtain (see appendix 3)

$$\frac{V_{QR}}{V_{PT}} \approx \frac{a}{2D} = \frac{R_D}{R_T + R_P} \qquad (5)$$

where a is the electrode diameter, D is the spacing from P to Q and R_D is a measure of the resistance between P and Q. The attenuation of the signal now depends not only on the electrode sizes (or resistances), as it did before, but on the ratio of the electrode sizes to the spacing between the ends of the break.

As before, the electrodes at Q and R do not appear in the expression because the receiver has a very large input resistance.

If the receiver *does* draw current then the expression is a little more complicated. Appendix 3 shows that the conduction paths in the rock can be modelled as shown in figure 6, with a schematic as shown in figure 7. (This location of R_D is not the only possible model, but others should be used with care. Not all these resistors relate explicitly to resistances in the ground.) The flow of current is more complex than the simple case of two electrodes – the resistance between the electrodes is clearly not exactly as we defined earlier, and R_D does not have a unique position in the model.

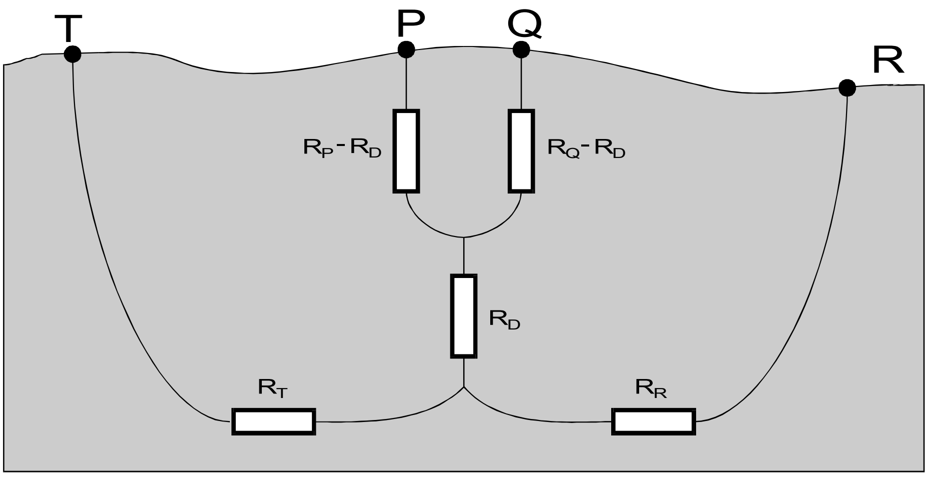


Figure 6 – One of several representations of the current paths for the array of figure 5. *R_T, R_P, R_Q, R_R, represent the electrode resistances, according to our particular definition. R_D represents the resistance due to the separation of P and Q. This representation is not unique and needs treating with care.*

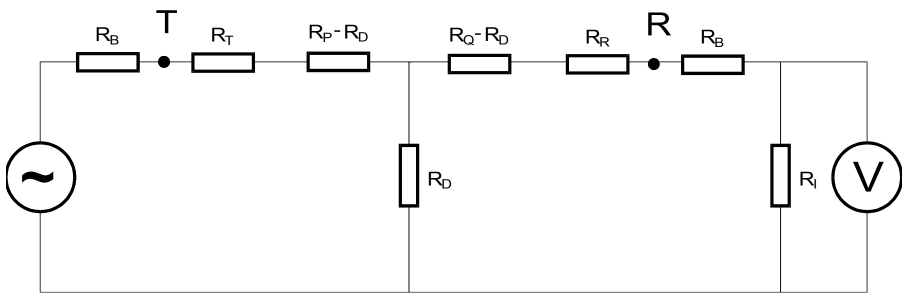


Figure 7 – A schematic version of figure 6.

Electrode / Body Resistance

Before applying the above model, we need to know likely values of R_b, R_T, R_R etc.

Metal Stakes

Suppose the *T* and *R* electrodes are metal stakes 250mm × 5mm diameter. If the rock has conductivity σ = 0.0001/Ωm then the resistance of each electrode, from (2), will be 33.7kΩ and the resistance between them will be double this, 67.5kΩ.

The resistance can be lowered by "watering" the electrodes with a solution of electrolytes. If we managed to saturate a hemisphere of earth of radius 250mm then we could expect the resistance between them to drop from 67.5kΩ to 12.7kΩ.

Human Body Resistance

The resistance between wet palms pressed on a conducting surface can be as low as 10kΩ. We could guess that a wet palm might, as an electrode, have roughly the same resistance as a 100mm diameter hemisphere, say 64kΩ with σ = 0.0001/Ωm.

Capacitive Coupling

Nigel Lovell's high-Z SWTs have been demonstrated to work when there is no body contact at all with the ground. This was put down to the capacitance between the body and ground having a low enough reactance at audio frequencies. Figure 8 shows the electric field lines associated with this capacitance. If the ground is conductive enough, then this essentially generates a "reflection" of the caver in the ground. The "virtual" electrode formed by the caver's reflection is large, and will have only a small resistance The capacitance of a sphere of radius 2m is around 56pF, with a reactance of 2.9MΩ at 1kHz. This is only an approximation to reality, of course, but is shows that, even without any body contact, we can expect the electrode resistance to be around 3MΩ. This is why, with Lovell's 10MΩ input resistance, his 'phones do not need a good earth.

A Short to Ground

If the "short" at *P* was a small nick in the wire, or a small twisted joint, then it can, perhaps, be modelled as a hemisphere 5mm in diameter, with a resistance of 637kΩ. In practice, if the nick was lying in a pool of water (say a hemisphere 250mm in diameter) then it would have a resistance of the same order of magnitude as the earth stakes we discussed earlier.

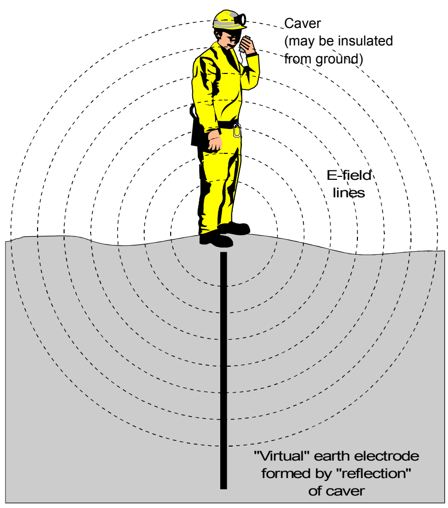


Figure 8 – A caver can form a large (capacitive) electrode without needing to make physical contact with the ground

Modelling SWT behaviour

Ignoring complications such as a break in the wire, we can use figure 3 to determine SWT performance in various circumstances. Our model is, of course, only an approximation – it assumes that the ground is homogenous, that the electrodes lie on the surface of an infinite half-plane, and that the spacing is much greater than their diameter (real or "virtual"). What is given here is, of necessity, a summary. A more complete discussion may appear at a future date.

Case 1 – High Resistance Receiver

The limiting case is where $R_i >> R_b + R_T$. There is little attenuation of the signal even if the human body is making poor ground contact (as in Nigel Lovell's design).

Short-circuit at P

If R_b is due to direct body contact then, from the figures given earlier ($R_b \approx 10k\Omega$, $R_T \approx 64k\Omega$ and $R_P \approx 13k\Omega$), we would expect the signal to be attenuated by 16.5dB. If the "short" were not in a pool of water, or if the Tx earth were improved, then the attenuation would be less. Clearly, if R_b is high, due to capacitive reactance, little signal will get through. But note that if $R_b \approx 3M\Omega$ then a value of $R_P \approx 500k\Omega$ will give a similar attenuation; so if the "short" is a "pin-prick" then some signal will still get through, even with a very poor Tx earth.

Case 2 – Low Resistance Receiver

The limiting case is $R_i << (R_b, R_T)$ and ultimately $R_i = 0$ in a current-sense configuration. (e.g. have $R_i = 100\Omega$ and use a gain of several hundred). Unlike the previous example, the signal attenuation now depends heavily on R_b, R_T even without a short at P. However, if we use a pseudo-current source (e.g. a high voltage in series with a high resistance) then changes in R_b, R_T can be made to have little effect.

Short-circuit at P

If the Tx current is more or less constant then it is shared between R_P and $R_b + R_T$ in proportion to their values, so the attenuation is similar to that in the first example. However, it is clearly now the receiver earth which must be improved to aid reception, not the transmitter earth, as it was previously.

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Further CREGJ Reading

Inductive and capacitive coupling have been covered in past journals, nos. 13, 17, 19, 25.

Equipment using an LF carrier for extra immunity to poor earth, and AGC action to compensate for attenuation caused, was reviewed in no. 27.

A current-driven system was described in no. 13.

Appendices

1: High Impedance Receiver

Using the "dualled" model with charge and capacity, as in Gibson, 1994, we can write the *applied* potential between T and P in Figure 4 as

$$\frac{Q}{4\pi\varepsilon}\left(\frac{1}{a_T}+\frac{1}{a_P}\right) \qquad \text{(A1)}$$

It will be somewhat simpler to do the calculations directly in terms of current so, for a half-plane (A1 was for a full-plane), we have

$$V_{PT} \approx \frac{-I_T}{2\pi\sigma}\left(\frac{1}{a_T}+\frac{1}{a_P}\right) \qquad \text{(A2)}$$

For the receiver we have

$$V_{PR} = \frac{I_T}{2\pi\sigma} \left(\left(\frac{1}{TP-a_P}-\frac{1}{a_P}\right)-\left(\frac{1}{TR-a_R}-\frac{1}{PR-a_R}\right)\right) \qquad \text{(A3)}$$
$$\approx \frac{-I_T}{2\pi\sigma}\frac{1}{a_P}$$

so we can write

$$\frac{V_{PR}}{V_{PT}} = \frac{R_P}{R_P+R_T} \qquad \text{(A4)}$$

where R_P, R_T are the resistance of the single electrodes P and T as discussed above.

2: Low Impedance Receiver

We define the current flowing *out* of the ground at each electrode as I_T, I_P, I_R; the hemispherical electrode radiuses as a_T, a_P, a_R, and the distances between the electrodes (on the surface of an infinite half-plane) as TP, PR and TR. The potential between the P and R electrodes, which is what will be measured by the receiver, is then given by

$$V_{PR} = \frac{1}{2\pi\sigma}\left\{\begin{aligned}&\left(\frac{I_T}{TP-a_P}+\frac{I_P}{a_P}+\frac{I_R}{PR-a_P}\right)- \\ &\left(\frac{I_T}{TR-a_R}+\frac{I_P}{PR-a_R}+\frac{I_R}{a_R}\right)\end{aligned}\right\} \qquad \text{(A5)}$$
$$\approx \frac{1}{2\pi\sigma}\left(\frac{I_P}{a_P}-\frac{I_R}{a_R}\right)$$

and similarly,

$$V_{PT} \approx \frac{1}{2\pi\sigma}\left(\frac{I_P}{a_P}-\frac{I_T}{a_T}\right) \qquad \text{(A6)}$$

We can now write these expressions using our *defined* terms for the resistance of the single electrodes, thus:

$$\left.\begin{aligned}V_{PT} &\approx I_P R_P - I_T R_T \\ V_{PR} &\approx I_P R_P - I_R R_R\end{aligned}\right\} \qquad \text{(A7)}$$

Now, noting that

$$I_T + I_P + I_R = 0 \qquad \text{(A8)}$$

and

$$V_{PR} = I_R R_i \qquad \text{(A9)}$$

we can proceed to solve the above equations if desired. The solution is tedious, but we can note that the equations describe, precisely, the configuration depicted in figure 3. The concern that the field lines could distort the current flow was unjustified.

3: Modelling a Wire Break

Proceeding in a similar fashion to the previous examples we can write

$$V_{QR} = \frac{1}{2\pi\sigma}\cdot \left\{\begin{aligned}&\left(\frac{I_T}{TQ-a_Q}+\frac{I_P}{PQ-a_Q}+\frac{I_Q}{a_Q}+\frac{I_R}{QR-a_Q}\right)- \\ &\left(\frac{I_T}{TR-a_R}+\frac{I_P}{PR-a_R}+\frac{I_Q}{QR-a_R}+\frac{I_R}{a_R}\right)\end{aligned}\right\} \qquad \text{(A10)}$$

We can write V_{PT} in a similar fashion. Now, if PQ is much smaller than the other distances; noting that $I_P = -I_T$ etc.; and writing PQ as D we obtain the simultaneous equations

$$\left.\begin{aligned}V_{QR} &\approx \frac{-I_T}{2\pi\sigma D} - I_R\left(R_Q+R_R\right) \\ V_{PT} &\approx \frac{-I_R}{2\pi\sigma D} - I_T(R_T+R_P)\end{aligned}\right\} \qquad \text{(A11)}$$

Writing

$$R_D = \frac{1}{2\pi\sigma D} \qquad \text{(A12)}$$

for convenience, we can re-arrange the terms:

$$\left.\begin{aligned}-V_{QR} &\approx \left(I_T+I_R\right)R_D + I_R\left(R_Q+R_R-R_D\right) \\ -V_{PT} &\approx \left(I_T+I_R\right)R_D + I_T\left(R_T+R_P-R_D\right)\end{aligned}\right\}$$

(A13) from which we can easily produce the diagrams of figures 6 and 7.

High-Resistance Receiver

If the receiver draws no current then the equations simplify to

$$\left.\begin{aligned}-V_{QR} &\approx I_T R_D \\ -V_{PT} &\approx I_T\left(R_T+R_P\right)\end{aligned}\right\} \qquad \text{(A14)}$$

from which we can produce (5) in the main text.

Wenner Array

As a check on the validity of (A10), let us suppose that $TQ = PR = QR = \alpha$ and $TR = PQ = 2\alpha$, which is the case for a Wenner array, so

$$\frac{V_{QR}}{V_{PT}} \approx \frac{-a}{2\alpha} \;\Rightarrow\; \frac{V_{QR}}{I_T} = \frac{1}{2\pi\sigma\alpha} \qquad \text{(A15)}$$

which is a standard result for ground conductivity as determined by a Wenner array experiment.

CREG

Building them Tough – Rugged PCs Investigated

*Think that caves and computers don't mix? **Mike Bedford** looks at some PCs which have got what it takes to withstand the cave environment.*

Introduction

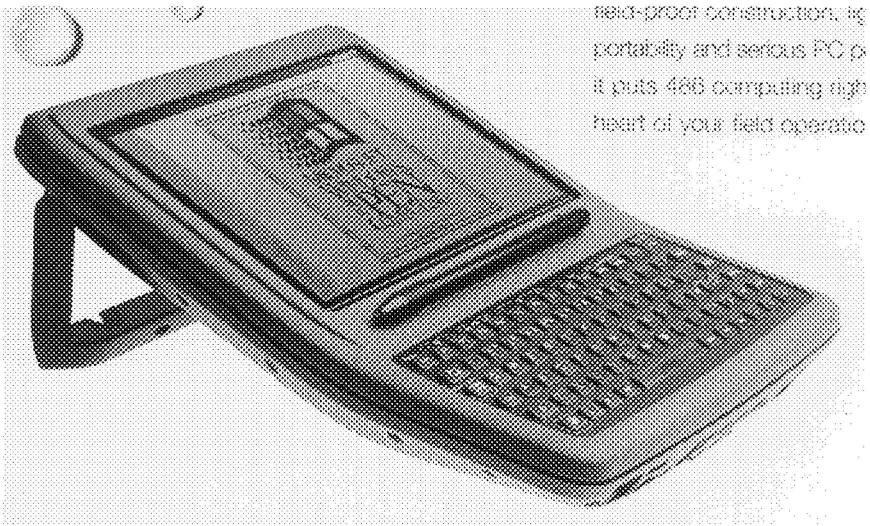
It was over four years ago that we last put rugged PCs under the spotlight (Bedford, 1993). And this certainly isn't a area in which things stand still so an airing of the subject is well overdue.

Four years ago, rugged PCs were few and far between, and were manufactured by a handful of specialist manufacturers. The computing specification was modest in the extreme – such was the price of ruggedness. Specifically, they used V20 or V25 processors (similar to the 8088 which was used in the original IBM PC), and had minute hard disks in the 4Mbyte region (actually they had solid state memory emulating a hard disk) They had very small CGA screens and were totally unsuitable for running modern Windows software. Yet the price was typically well over £1,000.

Today, there are many portable PCs boasting some degree of ruggedisation with specifications not dissimilar from standard business laptops. They are still produced mainly by specialist manufacturers although at least one mainstream supplier, Panasonic, has a rugged offering in the form of the CF-25. There is still a price premium associated with ruggedness, however, with prices varying from around £2,000 to almost £8,000 depending on the degree of ruggedness, the processing power and the expansion capabilities.

But it isn't only the supply of rugged portables which has changed over the course of four years – so has the attitude of cavers. In 1994 we covered rugged laptops as something of a novelty – any suggestions of real world caving benefits were somewhat contrived. An apparently obvious application was in recording survey data but four years ago many cave surveyors seemed to have a decidedly "neo-Luddite" outlook. OK, we still haven't got to the stage where every caver is going to need a trusty laptop with him on a sporting trip, but things are changing. Two articles in this *Journal* (Rabson, 1997; Sellers & Chamberlain, 1997) describe underground computing applications and the latter is surveying-related.

In this article we take a brief look at the rugged portable PC market in general and go on to take a rather more detailed look at a couple of units which may prove suitable for caving use.



The Husky FC-486

Roundup

Although they are still available, this article will not consider those hand-held units mentioned earlier which are based on the V20 or V25 processor. In reality, these are only suitable for simple data gathering applications and, in most cases, will be used in conjunction with custom-written software. Having excluded these PCs, we are left with about a dozen portable rugged computers on the UK market. All are capable of running Windows software and the processors range from the 25MHz 486SLC to the 166MHz Pentium. You'll notice that I haven't used the terms laptop or notebook since, in some cases, these descriptions would be inappropriate. Some of the units don't fold up, so clearly they can't be called notebooks, and some are really too heavy to be perched on your lap. In the main, the units in this latter category are intended for industrial data gathering and have extensive ISA and/or PCI expansion capabilities.

The manufacturers of all these dozen units make some claim that their products are ruggedised, but there are big differences in how well these units will stand up to harsh treatment. As a very minimum, these rugged PCs can withstand shock and vibration. Vibration won't be a major concern to cavers but shock certainly will – caving equipment needs to be protected against accidental drops. The US military standard Mil-Std 810E, method 516.4, procedure IV specifies that the unit must be capable of surviving drops from a height of 1.2m (so long as the largest dimension is less than 910mm) onto a steel plate over concrete. Furthermore, drops must be made onto all 26 faces, edges and corners of the unit under test. Less formally, you'll sometimes see a rather vague drop test figure such as "2m onto concrete" or "1m onto a carpeted floor". I would suggest that the Mil-Std 810E 516.4 IV is the minimum acceptable standard for caving use. Some manufacturers quote a drop test figure in addition to the military standard, to indicated that it exceeds this requirement, but a single drop test figure should be queried. It's pointless having a PC which can be dropped onto its base but not onto its corners or edges.

As an alternative to the drop test, some companies will specify shock in terms of acceleration or deceleration. This is measured in multiples of **g** – the acceleration due to the earth's gravity, not grammes. If a unit is dropped, it will always accelerate at **g**.

However, when it hits the floor, it will decelerate at a much higher rate, depending on the elasticity of the surface onto which it falls and of the unit itself. You'll see a wide range of figure quoted – from 15g to 100g. This is the basis of testing within the IEC 68-2-27 specification, which tends to be used by European manufacturers instead of Mil-Std 810E. Specifically, it requires that units must be able to sustain a 15g shock and this is applied using specially designed test equipment as opposed to dropping the unit. Unfortunately, it's not at all easy to relate these figures to dropping a unit and it's not easy to compare units for which different standards have been quoted.

The second area associated with ruggedness is waterproofing and this is obviously of major concern to cavers. Ironically, a few of the PCs which boast a good degree of ruggedisation have no protection against water whatsoever. Waterproofing and protection against the harmful ingress of solid objects or particles is specified as an IP number. The first figure relates to solid objects and the second to waterproofing. For caving use, we can forget about the lower specifications. All caving equipment must be protected to IP66, IP67 or IP68. The first 6 in each case indicates that the unit is "totally protected against the ingress of dust". 6 as a second digit indicates "protection against damage by jets of water (e.g. heavy seas)", 7 indicates "protection against damage by brief submersion to 1m" and 8 indicates "protection against damage by continuous submersion at a specified pressure". Specific test procedures are associated with the IP numbers but, as with shock, some manufacturers quote a less rigorously defined specification such as "drip proof" or "protected against jets of water".

When we take the caving requirements of shock resistance and waterproofing together, our list of a dozen rugged portables comes down to just two units. These are described in the following sections.

Husky FC-486

Husky is a British company with a reputation of building the ultimate in rugged portable computers. The FC-486 is certified to Mil-Std 810E 516.4 IV, although the data sheet additionally specifies that it can withstand a 2m drop onto concrete. Turning to waterproofing, the FC-486 boats IP67 – that is, it can be immersed in a metre of water for short periods of time. This was demonstrated at the March field meeting.

As always, there's a price to be paid for this degree of environmental protection, indeed the FC-486 is far behind standard business laptops in terms of its performance. The processor is a very modest 25MHz 486SLC which is Cyrix's version of the 486SX with a reduced width data bus. Optionally, a 50MHz 486SLC2 with numeric co-processor may be fitted. In common with earlier Husky computers, the FC-486 doesn't actually have a hard disk. Instead, it uses solid state memory to emulate the disk and, although this results in a much more resilient unit, this pseudo-disk is very small. The unit can be configured with up to 16Mbytes of memory and this must be shared between the disk and standard memory. However, PCMCIA cards can be used to provide additional flash memory. Similarly, the FC-486 doesn't have a floppy disk drive so applications would have to be installed via a serial port. Data would be uploaded to a desktop PC via the same mechanism.

Turning to styling and ergonomics, the FC-486 is rather like a miniature notebook except for the fact that it's always open. At 2.3kg, it's reasonably light and it can easily be used whilst standing. The screen is monochrome (an advantage in bright sunlight) and is equipped with an optional back light. The screen is touch sensitive with a pen being used as the pointing device. The pen is of an active design, however (i.e. you can't use your finger) so if the pen gets lost (there is no tether) then you've got problems. And finally, the FC-486 is very frugal in its power requirements – it will run for 10 hours on a single charge.

Telxon X-C6000

In terms of its appearance, the Telxon X-C6000 (manufactured by parent company Itronix in the USA) is more like a standard notebook PC, albeit a rather small one with a 8.2" screen. Despite its size, though, it's not particularly light at 2.9kg – clearly this is an inevitable result of ruggedisation. The unit is coated with what appears to be non-slip rubber. However I'm told that, unlike rubber, the coating will absorb the energy from a shock so the unit will only suffer a single shock if it's dropped – i.e. it won't bounce. Whilst on the subject of its physical characteristics, it's appropriate to point out that the unit has a very good feel to it, both in use and whilst it's being carried.

The X-C6000 is, supposedly, certified to Mil-Std. 810E 516.4 IV, although the data sheet then goes on to mention a 1m drop test figure which is a less stringent requirement that that required by the military standard. As a true notebook, the hinge may be a cause for concern, but Telxon were at pains to point out that this its isn't a weak spot. Telxon salesmen demonstrate this by placing a unit on the floor, half opened in an "A" configuration, and then standing on the apex.

Unfortunately, an IP rating is not quoted although the manufacturers indicate that the unit is "protected against jets of water". This sounds like IP66, although the company tell me that they have demonstrated a unit working whilst it was inside a washing machine – more like IP67 conditions. From the shock and waterproof figures, the X-C6000 is less obviously cave-proof than the FC-486 but the various inconsistencies mean that we can't be sure about its suitability without asking further questions.

As one of the more rugged PCs available, you might reasonably expect the processing power to be compromised and you'd be correct. The unit is currently available with a 50MHz 486SLC2 although faster processors are in the pipeline. The unit can be supplied with a maximum of 16Mbytes of RAM although, unlike the Husky, it can all be used as "real memory" since it doesn't have to be shared with the solid state disk. The X-C6000 has a real hard disk although at just 260Mbytes it's not exactly huge.

Like the FC-486, the Telxon PC has a monochrome screen (colour will be available soon) which is also touch-sensitive. Unlike the Husky, however, you don't have to use a special active pen, you can use virtually any pointing device including your fingers. Actually, although the unit supplied for evaluation did have a touch screen, this is optional and an alternative pointing device in the form of a nipple is also provided.

Sources

The Husky FC-486 costs £4,328 + VAT for a model with the 25MHz processor and the maximum 16Mbytes of memory. Contact Husky Computers Ltd. on 01203 604040. In the USA, contact Husky Computers Inc. on (813) 530 4141. Husky also have subsidiaries in France and Germany.

The Telxon X-C6000 costs £3,900 + VAT for the specification discussed in the article. Contact Telxon Limited on 01202 785300. Telxon International in Belgium is on +32 2 663 1500 and Itronix (the US parent company) is on +1 800 555 4104.

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LF Utility Stations

A list of transmitters operating in the LF spectrum, compiled by ***David Gibson*** *from information supplied by Joerg Klingenfuss of Klingenfuss Publications, and others.*

This table of LF utility stations has been complied from a list entitled "1996 Super Frequency List on CD-ROM", published by the German company *Klingenfuss Publications* (see box). It is reproduced here by permission. Thanks are due also to John Hey and Nick Potter for the information about the Racal Decca navigation beacons.

Standard Time Stations

I have added the 16kHz GBR beacon to the table, and if anyone has any information about it, I would be interested to know. (Is it still operational?)

I have found the 60kHz MSF beacon useful for measuring the performance of cave radio receivers. It has a broadcast power of 27kW. On a 'scope you can easily see the one second "pips", the slow-rate data and the burst of fast data just before the minute signals. The data format was published a while ago in *Wireless World*.

Chain (& ident.)	Purple Slave (5×f)	Stations (kHz) Master (6×f)	Red Slave (8×f)	Green Slave (9×f)
SW British (1B)	70.2333	84.2800	112.3733	126.4200
Northumbrian (9A)	70.3792	84.4550	112.6067	126.6825
North British (3B)	70.5375	84.6450	112.8600	126.9675
English (5B)	70.8333	85.0000	113.3333	127.5000
N. Scottish (6C)	70.9875	85.1580	113.5800	127.7775
Hebridean (8E)	71.3625	85.6350	114.1800	128.4525

Racal Decca Navigation Beacons around the UK.

Racal-Decca Nav. Beacons

I have also added the these 1-2kW beacons, which form part of a world-wide hyperbolic navigation system. It is unlikely that, as cavers, we will need to know details such as location, data format, etc., but this information is available if required.

Abbreviations Used

A * in the "details" column of the table below denotes that the frequency is listed in Klingenfuss's "formerly active frequencies" list, and only the approximate frequency is shown. A list of abbreviations is given at the end of the article.

Klingenfuss Publications

The data given here is an extract from the "Utility Stations 1996" and "Formerly Active Stations" sections of Klingenfuss Publications' "Super Frequency List on CD-ROM".

Only a small amount of the CD-ROM data is concerned with LF stations. The 1997 issue contains 11,500 entries for broadcast stations world-wide, 13,800 utility stations, and 14,100 formerly active stations from 0-30MHz. You can search for specific frequencies, countries, stations, languages, call-signs and times. *Klingenfuss* sell a number of radio guides, both on paper and CD-ROM, as well as radio hardware such as the Wavecom W41PC DSP Data Decoder Card. Their publicity information states: "Klingenfuss Publications are compiled from our own monitoring of the radio spectrum according to an exact shift schedule which covers all frequencies at all times during a four-monthly turn. Contrary to other "authors", we do not "rely" on any type of foreign material from so-called "reliable sources" ... and we do not plagiarise! All publications are compiled by means of a self-programmed computer-controlled data bank and word processor system.

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Table of LF Utility Stations *(continued on next page)*

Frequency (kHz)	Callsign	Station	Country	Modulation	Details
16.0	GBR	Rugby TS			??
17	UMS	Moscow, RU	URS	CW	*
18.3		UNID teleprinter		RTTY	200bd
23.4	DHO 38	GN West Rhauderfehn	D	CW	
23.4		UNID teleprinter		RTTY	50bd
24	NBA	USN Balboa	PNR	CW	*
25	UNW 3	Kaliningrad TS, RU	URS	CW	*
25	UPD 8	Murmansk TS, RU	URS	CW	*
25	UTR3	Gor'kiy TS, RU	URS	CW	*
38.0	SHR	SN Ruda	S	CW	
50.0	RTZ	Irkutsk TS, SE	RUS	CW	
51.9		UNID teleprinter		RTTY	50bd
53.0		UNID teleprinter		RTTY	75bd
53.6	RTO	Moscow Meteo	RUS	FAX	
55	DCF 55	VWD Frankfurt	D	RTTY	*
60.0	MSF	Rugby TS	G	CW	
60.0	WWVB	Fort Collins TS, CO	USA	CW	
62	GIZ 20	RN London	G	CW	*
62.6		UNID teleprinter		RTTY	200bd
63	FTA	FN Paris	F	RTTY	*
63.9		UNID teleprinter		RTTY	200bd
65.8		UNID teleprinter		RTTY	200bd
67	RBU	Moscow TS, RU	URS	CW	*
69.0		UNID teleprinter		RTTY	50bd
69.1	RCK	Novosibirsk Meteo, SW	RUS	FAX	
70 - 71		*Racal Decca* Purple			
73.3	MTO 21	RN Crimond	G	RTTY	ITA2 50bd

Table continued ...

75.0	HBG	Nyon TS	SUI	CW	
77.5	DCF 77	Mainflingen TS	D	CW	
82.8	MKL	RAF Edinburgh	G	CW	
83	GYB	RN London	G	CW	*
84	FTA 83	Paris TS	F	CW	*
84 - 86		*Racal Decca* Master			
85.6	IDQ	IN Rome	I	CW	
90-110		*Loran* wideband			
111	DCF 30	DPA Hamburg	D	RTTY	*
111.3	SOA 211	Warsaw Meteo	POL	RTTY	ITA2 50bd
111.8	OLT 21	Prague Meteo	CZE	FAX	
112 - 114		*Racal Decca* Red			
117.4	DCF 37	Offenbach Meteo	D	FAX	
118.5	IDQ	IN Rome	I	CW	
119.5	SXA	GN Piraeus	GRC	CW	
120	SAY 2	Norrköping Meteo	S	FAX	*
120.9		UNID teleprinter		RTTY	75bd
122.3	OUA	DM Stevns	DNK	CW	
122.5	CFH	CF Halifax, NS	CAN	FAX/RTTY	ITA2 75bd
122.7	JMC	Tokyo Meteo	J	CW	
124	CKN	CF Victoria, BC	CAN	CW	*
124	DCF 42	PIAB Bonn	D	RTTY	*
125	OLT 3	Prague Meteo	TCH	FAX	*
125	SLZ	SAF Kristinehamn	S	CW/FAX	*
125.9		UNID teleprinter		RTTY	75bd
126 - 129		*Racal Decca* Green			
129	DCF 45	VWD Frankfurt	D	RTTY	*
129.1	DCF 49	BMPT Bonn	D	RTTY	IRA 200b
129.5	SOA 212	Warsaw Meteo	POL	RTTY	ITA2 50bd
134.2	DCF 54	Offenbach Meteo	D	FAX	
139	DCF 39	DPA Frankfurt	D	FAX	*
139.0	TBA	TN Ankara	TUR	CW	
140	DCF 60	EPD/SID Frankfurt	D	RTTY	*
144.5	RCG	Moscow Meteo	RUS	FAX	
146	Y7A 20	MFA Berlin	DDR	RTTY	*
147	DDH 47	Hamburg Meteo	D	CW	*
147.3	DDH 47	Hamburg Meteo	D	RTTY	ITA2 50bd
147.8		UNID teleprinter		RTTY	100bd

The following abbreviations are taken from chapter 24 of the Klingenfuss “1996 Guide To Utility Radio Stations”

BC British Columbia / Broadcast
CAN Canada
CF Canadian Forces
CO Colorado
CW Continuous wave (Morse code)
CZE Czech Republic
D Germany
DDR German Undemocratic Republic
DM Danish Marine
DNK Denmark
DPA Deutsche Presse-Agentur
EPD Evangelischer Pressedienst
F Fisheries / France / French
FN French Navy
G German / United Kingdom
GN German or Greek Navy
GRC Greece
I Italian / Italy
IN Indiana / Indian or Indonesian or Irish or Israel or Italian Navy
J Japan
MFA Ministry of Foreign Affairs
NS Nova Scotia
PNR Panama
POL Poland
RN Royal Navy
RTTY Radioteletype
RU Russia (W of 60 E)
RUS Russian Federation (areas east of 60 E are additionally specified by SE and SW)
S South / Spanish / Sweden
SAF Spanish Air Force
SE Russian Federation (Eastern Siberia 100 E - 140 E)
SID Sport-Informationsdienst
SN Spanish or Swedish Navy
SUI Switzerland
SW Russian Federation (Western Siberia 60 E - 100 E)
TCH Czechoslovakia
TS Time Signals and/or Standard Frequency Station
TUR Turkey
UNID Unidentified
URS Union of Soviet Socialist Republics (up to 20 DEC 1991)
USA United States Army / United States of America (except Alaska and Hawaii)
USN United States Navy
VWD Vereinigte Wirtschaftsdienste

CREG

Constructing the 3496Hz “D-Q” Beacon Receiver

(continued from page 7)

Semiconductors

D1, D6	Red LED
D2	IN4001
D3, D4, D5	IN4148 or IN914
L1	Green LED
U0	LF356
U1, U2	LF412
U3, U4, U8	LF442
U5	4016
U6	4060
U7	4070
U9	LM386
U10	78L12 (could be ECG-950)
V1	ECG-614 or 1N5470A (Varactor diode about 33pF at 4V reverse)
X1	3.579545MHz (Colour burst crystal)

Miscellaneous

Alarm	buzzer
B1, B3	BNC (B1 insulated from case)
Batt1, Batt2	9V
Header1	8 pin header
Header2	16 pin header
Loop	18.25”
P1, P2	1/8” phone jack
S1	DPST
S2, S4	SPST
S3	SPDT

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