10 Low Frequencies: Below 1MHz

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The 136kHz band, 135.7kHz - 137.8kHz was introduced in January 1998 and is unique in being in the LF frequency range (Low Frequency, defined as 30kHz - 300kHz).

In February 2007, the licensing authority, Ofcom. began inviting applications for a Notice of Variation for UK amateurs to operate for experimental and research purposes in the range 501.0kHz - 504.0kHz [1] in the MF range (the Medium Frequency range, defined as 300kHz - 3MHz, and also including the 160m band). The issued NoVs have been renewed periodically by Ofcom; current NoVs are valid until February 2012.

Both 136kHz and "500kHz" bands have unique characteristics and are different to all the higher frequency amateur bands. Propagation on 136kHz and 500kHz is very different to the HF bands (for more on this, see the chapter on propagation).

Due to the narrow bandwidth available at 136kHz (a total of only 2.1kHz), the low radiated power level permitted (1W ERP) and the high noise levels present on this band, several specialised techniques [2, 3] have been developed for 136kHz DX operation, alongside familiar CW for shorter-distance contacts. Some LF DX modes are described in the Digital Communications chapter.

The 500kHz band has a somewhat lower noise level, and the radiated power limit has recently been increased to 10W ERP, but offers its own challenges, particularly the very deep fading that occurs at intermediate and long distances. The majority of 500kHz operation has also been in CW, although digital modes have seen increasing use. In the future, it can be expected that both bands will see the development of new operating modes and techniques to achieve communication under the often very marginal conditions.

RECEIVERS FOR 136kHz AND 500kHz

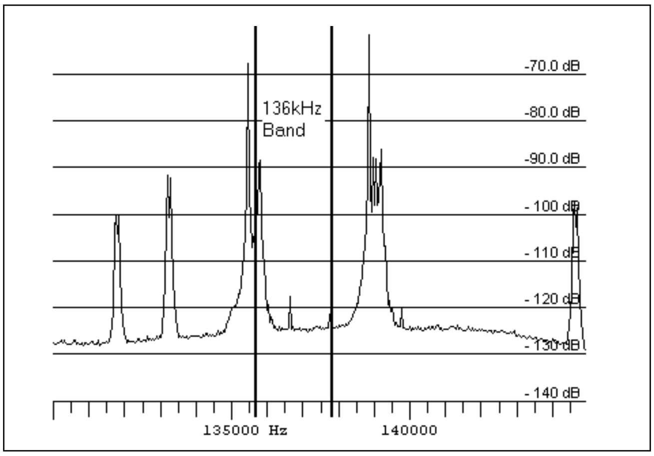
The majority of stations currently use commercially available receivers. Many amateur HF receivers and transceivers include general coverage that extends to 500kHz and 136kHz, and in many cases these can be successfully pressed into service; however, unlike HF reception, where reasonable results are often achieved simply by connecting a ‘random’ wire antenna to the receiver input, successful reception on these bands is a bit more difficult, for a number of reasons. First, there is usually a very large mismatch between the impedance of a wire antenna at these frequencies and the typically 50-ohm receiver input impedance, which leads to a large reduction in signal level at the receiver input. This is often exacerbated by degraded receiver sensitivity at 500kHz and below, particularly in amateur-type equipment. Secondly, amateurs with their relatively tiny radiated signals share the spectrum with vastly more powerful broadcast and utility signals; unless effective filtering is provided, this results in severe problems with overloading at the receiver front end. Fortunately, very satisfactory results can usually be achieved by using quite simple antenna matching, preamplifier, and preselector arrangements, as will be seen later in this section.

LF and MF Receiver Requirements

Requirements for LF receivers depend on the type of operation that is envisaged. Adequate sensitivity is obviously required; the internal receiver noise level should be well below the natural band noise at all times. As a guide, similar figures to those for HF receivers (a few tenths of a microvolt in a CW bandwidth) will suffice. If a large, transmitting-type antenna is used, the signal level will be high enough to allow a receiver with considerably poorer sensitivity to be used. Small loop and whip receiving antennas usually require a dedicated low noise preamplifier (see section on receiving antennas and noise reduction).

A widely used operating mode is CW. Because of the narrow bandwidth available (136kHz is only 2.1kHz wide, and 500kHz is 3kHz wide) a CW filter is almost essential. A 500Hz bandwidth is adequate, but 250Hz or narrower bandwidths can be used to advantage. On 136kHz, there are several strong utility signals just outside (and sometimes inside) the amateur band (**Fig 10.1**), and so good filter shape factor is important since the utility signals can be 60dB or more above the level of readable amateur signals, with a frequency separation of less than 1kHz. For specialised extremely narrow-bandwidth modes such as QRSS (see the chapters on Morse and Digital Communications), selectivity is also provided at audio frequencies using DSP techniques in a personal computer, but good basic receiver selectivity is still desirable to prevent strong out-of-band signals entering the audio stages.

The spectrum around 500kHz was once heavily used for maritime communications, but now non-amateur signals close to the 501kHz - 504kHz range are rare (Fig 10.1), making selectivity less critical.



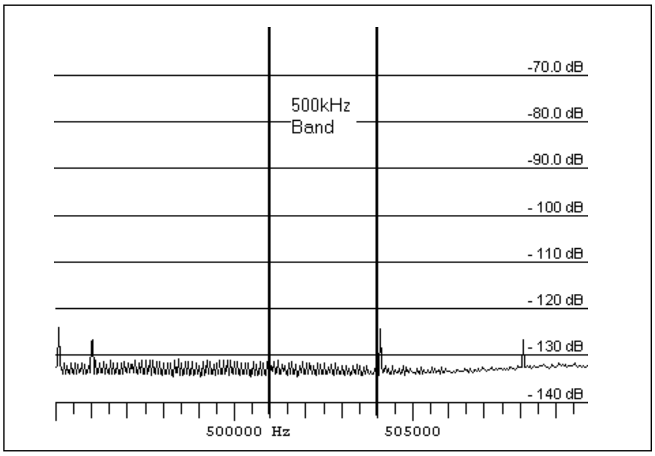


Fig 10.1: Spectrum in the vicinity of (left) 136kHz, and (right) 500kHz (vertical scale is field strength in dBV/m)

Frequency stability requirements depend on the operating mode. For CW operation, maintaining frequency within 100Hz during a contact is not usually a problem. Narrow band modes such as QRSS [2, 3] require better stability, and frequency setting accuracy. For the popular QRSS3 operating speed (widely used for 136kHz DX contacts around Europe), an initial setting accuracy, and drift of perhaps 10 - 20Hz per hour is acceptable if somewhat irritating. This level of stability can often be achieved by older receivers with mechanically tuned VFOs, provided the receiver is allowed to reach thermal equilibrium, and some means of calibrating the receiver frequency is available. These difficulties are eliminated in fully synthesised receivers, which generally exhibit a setting accuracy within 1 or 2Hz and drift a fraction of a Hertz over an extended period. This degree of frequency stability is adequate for the vast majority of applications, including the reception of inter-continental 136kHz beacon signals using QRSS30, QRSS60 and QRSS120 speeds, with bandwidths of as little as 0.01Hz, along with narrow-band, weak-signal digital modes.

For some specialised narrow band operating modes, a higher order of stability is required. This has been achieved by using high stability synthesiser reference frequency sources, such as TCXOs, OCXOs, and even atomic clock or GPS-derived frequency standards.

Amateur Receivers and Transceivers at LF

Many amateur HF receivers and transceivers can tune to 136kHz and below, and since they are already available in many shacks, probably a majority of 136kHz and 500kHz amateur stations use receivers of this type. All modern equipment is fully synthesised, so frequency accuracy and stability are good. Older receivers using multiple crystal oscillators or an interpolating VFO have relatively poor stability.

Modern crystal or mechanical CW filters have excellent shape factors, giving good rejection of strong adjacent signals. In some receivers, multiple filters can be cascaded, giving further enhanced selectivity.

Since reception at frequencies below 1.8MHz is generally included as an afterthought, manufacturers rarely specify sensitivity of amateur receivers at 136kHz and 500kHz. Unfortunately it can often be poor. There is little relation between the HF performance, cost, or sophistication of a particular model, and the sensitivity at lower frequencies. Therefore it may well be that older, cheaper models perform better at LF than their newer successors.

Few laboratory-quality sensitivity measurements are available for the LF and MF sensitivity of amateur transceivers and receivers, but the following lists some models which have been used successfully as LF receivers.

Classified as "good" are the Kenwood TS-850 (probably the most popular HF rig with 136kHz operators, but with reduced sensitivity at 500kHz - **Fig 10.2**), TS-440 and Yaesu FT-990 transceivers. Receivers include the AOR AR-7030, Icom R75, JRC NRD-345, NRD-525 and NRD-545, and Yaesu FRG-100.

Classified as "adequate" are the Icom IC-706, IC-718, IC-756Pro, IC-761, IC-765, and IC-781, Kenwood TS-140, TS-870 Yaesu FT-817 and FT-1000MP. Equipment classified as "adequate" requires either a large antenna and/or an external preamplifier to achieve adequate sensitivity. The IC-718 is fairly typical in this respect, requiring 1 microvolt at 136kHz to achieve 10dB SNR in a 250Hz bandwidth, a figure about 20dB poorer than it achieves in the HF bands.

The reason for poor sensitivity lies within the receiver front-end design. The inter-stage coupling components, in particular the first mixer input transformer, are optimised for operation at HF, and often have high losses at LF, reducing the signal level. Internally generated synthesiser noise may also be higher at LF. The front end filter used when the receiver is tuned to 500kHz or below is normally a simple low-pass filter with a cut-off frequency of 1 - 2MHz, often including an attenuator pad to reduce overloading due to medium wave broadcast signals; this further reduces sensitivity, without eliminating the broadcast signals. Some LF operators have improved receiver performance substantially by replacing the mixer input transformer with one having extended low frequency response [4]; this component must be carefully designed if receiver HF performance is to be maintained. A simpler and more common approach is to use an external preamplifier, and provide additional signal frequency selectivity, as described later in this section.

Commercial Equipment for LF

Many professional communications receivers made by such firms as Racal, Plessey, Harris, Collins, Eddystone, Rohde & Schwarz and others include coverage of the LF and MF spectrum, and surplus prices are often competitive with the amateur-type equipment discussed above. Ex-professional equipment is usually fully specified at LF and MF frequencies, so sensitivity and dynamic range are usually good at 136kHz and 500kHz. Fully synthesised professional receivers often have precision reference oscillators with excellent stability; they also usually have inputs for an external frequency reference. These features are not often found on amateur-type equipment, making them attractive if the more specialised LF communications modes are to be explored. A drawback is that affordable examples are usually fairly old, so servicing and repairs may be required from time to time. Also, they have a rather Spartan feel, with few of the ‘bells and whistles’ operator facilities found on modern amateur rigs. The Racal RA1792 (**Fig 10.2**) has been popular with UK amateurs on 136kHz and 500kHz. The older RA1772 also performs well.

A number of vintage receivers, including the HRO, Marconi CR100, AR88LF, cover 136kHz and 500kHz. Also, valve-era equipment designed for marine service often includes LF and MF coverage. A few amateurs have used vintage receivers for 136kHz operation. The antenna input circuit of this type of equipment is generally designed to be operated in the lower frequency ranges using an un-tuned wire antenna, and usually gives good sensitivity at 136kHz without requiring additional antenna tuners or preamplifiers. The major disadvantage of most vintage receivers is that their single-pole crystal filters have poor skirt selectivity compared to modern IF filters. This results in strong utility signals several kilohertz from the receive frequency reaching the IF and detector stages of the receiver,

Fig 10.2: Racal RA1792 (top), and RA1772 perform well on LF

causing blocking and heterodyne whistles which swamp the weak amateur signals. Unmodified vintage receivers are therefore usually poor performers at 136kHz, although for the experimentally minded, the addition of a modern IF filter and product detector could result in an effective LF CW receiver. As noted above, selectivity is less critical for 500kHz operation, and vintage receivers can perform quite well. An un-modified wartime HRO receiver has been used at M0BMU for 500kHz CW operation, with quite satisfactory results.

Selective level meters (SLMs), also called selective measuring sets or selective voltmeters, are instruments designed for measuring signal levels in the now-obsolete frequency division multiplex telephone systems; consequently, they are sometimes available surplus at low cost. SLMs are designed for precision measurement of signals down to sub-microvolt levels; their frequency range extends from a few kilohertz into the MF or HF range, so can make effective LF and MF receivers. Well-known manufacturers are Hewlett-Packard (HP3625) and the German companies Wandel and Golterman (the SPM- *selektiver pegelmesser* series; **Fig 10.3**) and Siemens.

SLMs are not purposely designed as receivers and do not have many normal receiver features, such as AGC and selectable operating modes, or sometimes even an audio output. Filter bandwidths are designed for telephony systems and are not always suited for amateur radio operating modes. Normally the 'CW pitch' is fixed at around 2kHz, so they are not well suited to CW operating, although this presents no obstacle for 'sound card' operating modes. The area where SLMs excel is in signal measurement; they have been used by a number of amateurs for 136kHz and 500kHz field strength measurements (see LF Measurements and Instrumentation section). They are often available with a tracking level generator, which is very useful for measurements on filters, or bridge-type impedance measurements.

Software-defined Radio Receivers

Software defined radio (SDR) is now becoming part of the mainstream of amateur radio, with both home constructed and commercially produced SDR hardware and software now widely available, see the chapter on software-defined radio. PC-based spectrogram software has been used for several years in conjunction with conventional receivers for the 'visual' LF/MF operating modes such as QRSS and narrow-bandwidth data modes; SDR is the natural extension of this trend.

Homebrew amateur SDR projects are most commonly based on PC-based digital signal processing software, using the PC sound card for A/D conversion of the incoming signal. Since the sound card is usually limited to 48kHz sample rate, the maximum signal frequency that can be handled by the sound card input is 24kHz. This allows direct reception of signals in the VLF range (see VLF section below), but for amateur band use, some form of external down conversion is required. This generally takes the form of an I/Q down converter, with in-phase and quadrature outputs feeding the left and right stereo inputs of the sound card.

The I/Q signal format permits image rejection to be performed by the SDR software, and also extends the bandwidth that can be processed by the sound card to 48kHz maximum; this is ample to cover the narrow amateur 136kHz and 500kHz bands with fixed, crystal-controlled conversion frequencies. All required tuning, filtering and demodulation functions are then performed in the digital domain by the SDR software. Several suites of SDR software have been made available free of charge for amateur use [5, 6, 7] that are suitable for use with I/Q down converters. This results in a very simple yet capable amateur band receiver; modifications to the well-known KB9YIG 'SoftRock' SDR receiver kits to permit 136kHz and 500kHz reception are described later in this section.

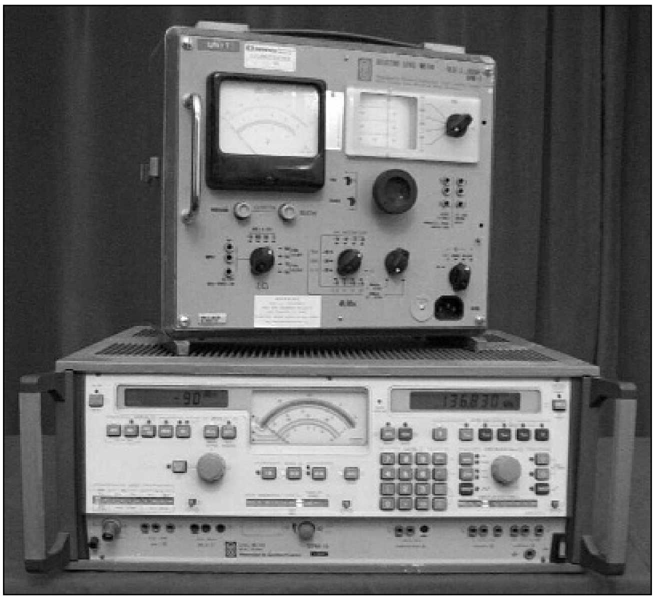


Fig 10.3: SPM-19 (bottom), and portable SPM-3 (top) selective level meters can be used as LF receivers

General coverage, direct-digitising SDR receivers are also becoming commercially available to amateurs at reasonable prices. Several amateur stations are successfully using the RFSpace Inc. SDR-IQ [8] and the Perseus SDR receiver [9] for LF and MF reception. Both these receivers are supplied with their own native SDR software, but can also be used directly with popular spectrogram software such as DL4YHF's *Spectrum Lab* [6] and *Winrad* [7] in order to generate high resolution spectrograms for 'visual modes' operation.

RECEIVE ANTENNA TUNING

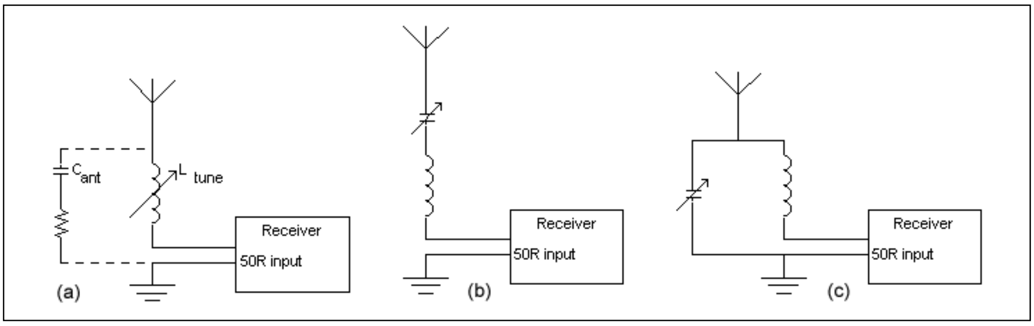
The impedance of a typical long-wire antenna at LF or MF can be modelled as a series resistor and capacitor. Taking the example in the Transmitting Antennas section of a typical long-wire antenna, the capacitance might be 287pF in series with 40 ohms at 137kHz. At 500kHz, the capacitance will be almost the same, but the resistance could be lower, perhaps 20 ohms. Assuming receiver input impedance of 50 ohms, the SWR at the feed point of the antenna is about 8200:1 at 137kHz! This mis-match results in an unacceptable signal loss of about 32dB. The loss due to mismatch at 500kHz is less severe, but still more than 20dB.

Most of the loss is caused by the capacitive reactance; signal levels can be greatly improved by resonating the antenna at the operating frequency with a series inductance. In the example above, the antenna with resonating inductance form a tuned circuit with Q around 40 and bandwidth of only a few kilohertz, which is very effective in filtering out powerful broadcast band signals.

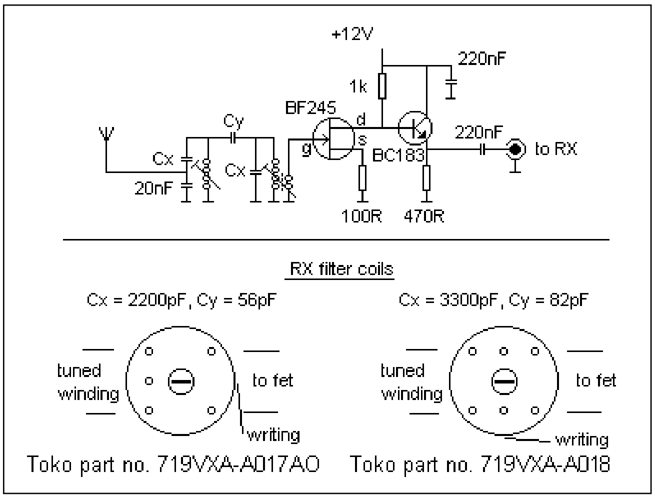
The practical effect of resonating the antenna is dramatic, normally with a long-wire antenna connected directly to the receiver, the only signals heard in the 136kHz range are numerous intermods. With the antenna tuned, these disappear and the band noise is audible above the noise floor of reasonably sensitive receivers. Attempts to receive signals at 500kHz with un-tuned wire antennas are more successful at locations where broadcast signal levels are fairly low, but an antenna tuner still yields substantial improvements.

Typical circuits used to tune wire antennas for reception are shown in **Fig 10.4**. **Fig 10.4(a)** is a simple series inductor; the value required is:

Fig 10.4: Receive antenna tuning circuits



(below) Fig 10.5: G3YXM's 136kHz preamp



$$L_{tune} = \left(\frac{1}{2\pi f\sqrt{C_{ant}}}\right)^2$$

with L_{tune} in henries, C_{ant} in farads and f in hertz. A useful rule of thumb is that the antenna capacitance C_{ant} will be roughly 6pF for each metre of wire, typically L_{tune} of a few millihenries will be required for 136kHz and a few hundred microhenries at 500kHz. Because of the high Q the inductance must be adjustable; this can be done using the same techniques as for transmitting antennas, or a slug-tuned coil can be used. It is often more convenient to use a fixed inductor, and adjust to resonance using a variable capacitor, as shown in **Fig 10.4(b)**. This can be a broadcast-type variable, with both sections paralleled to give about 1000pF maximum.

A higher tuning inductance is required to make up for the overall reduction in capacitance. The shunt-tuned circuit of **Fig 10.4(c)** has the convenience of one side of the tuning capacitor being grounded. The impedance match will not be quite as good, although normally perfectly adequate.

RECEIVING PREAMPS

To overcome reduced sensitivity at lower frequencies, many amateur-type receivers require a preamplifier. Because of the strong broadcast signals in the LF and MF frequency ranges, it is important that adequate selectivity is provided at the signal frequency. To obtain good S/N ratio, it is also necessary to pay attention to impedance matching between antenna and preamplifier.

A useful and well-tried design for 136kHz due to G3YXM is shown in **Fig 10.5**, and is available as a kit with PCB from G0MRF [10]. It incorporates a double-tuned input filter which provides a bandwidth of around 3kHz centred on the amateur band. The preamp is designed for 50-ohm input impedance, so antenna matching as described in the previous section will normally be required.

The LF antenna tuner/preamp circuit of **Fig. 10.6** combines the antenna matching, filtering and preamplifier functions. It is quite flexible and can be used with a wide range of long-wire and loop antenna elements. It can easily be modified for other frequencies, including 500kHz. It has been used successfully with an IC-718 transceiver, which has fairly poor sensitivity at 136kHz. The preamp is a compound follower, with a high-impedance JFET input, and a bipolar output to drive a low impedance load with low distortion. The gain of the follower is about unity, but the high Q, peaked low-pass filter input circuit provides voltage gain, and also gives substantial attenuation of unwanted broadcast signals at higher frequencies.

The gain of the circuit depends on the type of antenna element used, of the order of 10dB with a long wire element and 30 - 40dB with a loop element. The 2.2mH inductors are the type wound on small ferrite bobbins with radial leads, and have a Q around 80 at 136kHz; other types of inductor with similar or greater Q could also be used. For wire antennas, C_{in} should be in the range 600pF - 5000pF, with large values giving a reduced signal level with longer wires, and smaller values suiting short wire antennas.

The antenna can be fed with coaxial cable, in which case the distributed capacitance of the coax (about 100pF/m for 50-ohm cable) makes up part or all of C_{in}. This allows the receiving antenna to be located remote from the shack, which is often useful in reducing interference pick-up. This circuit has given good results with wire antennas ranging from a 5m vertical

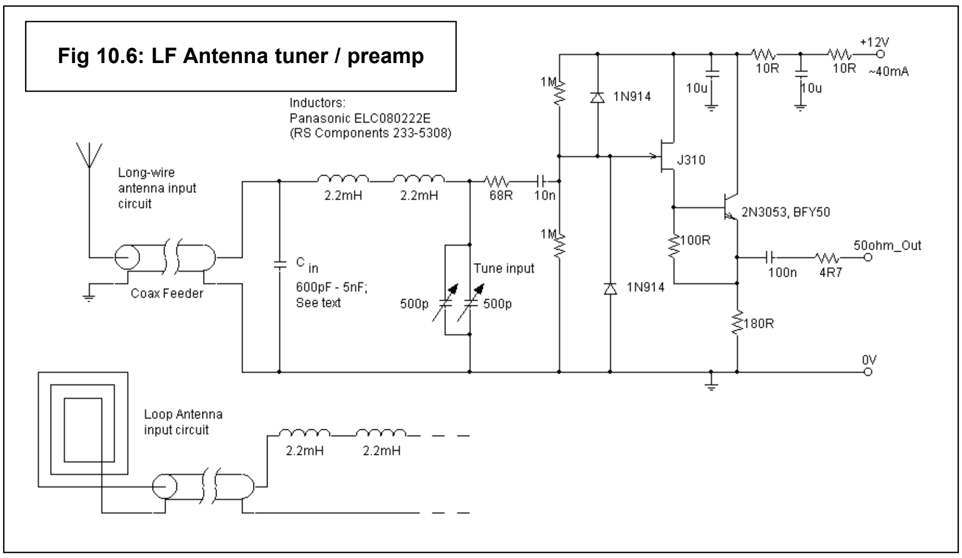


Fig 10.6: LF Antenna tuner / preamp

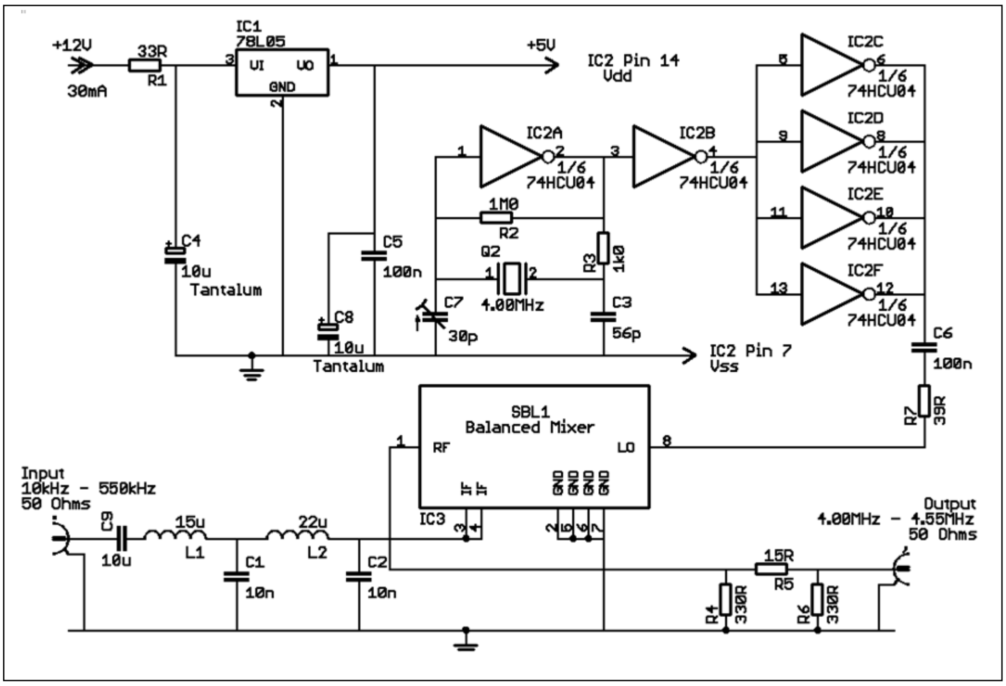


Fig 10.7: LF/MF - HF converter

whip to a 55m long wire. For loop antennas, C_{in} is omitted. Satisfactory sensitivity was obtained using a $1m^2$ loop with 10 turns of $1mm^2$ insulated wire, and also with larger single-turn loops with area around $10m^2$ (see the Receiving Antennas section). Loop antennas can also be fed via quite long lengths of coax cable. The frequency range of the input circuit can easily be extended by using different values of inductance in the input tuning circuit; for example, the circuit was used to receive the 17.2kHz VLF broadcasts from SAQ by using 200mH inductance, and has also been used at 500kHz with 330µH inductance.

CONVERTERS

For HF receivers without coverage of 136kHz, or those that have very poor performance at LF, a converter can offer excellent 136kHz reception. A number of LF/VLF - HF converters have been manufactured in the past (Datong VLF converter, Palomar VLF-S, VLF-A), although these have mainly been aimed at broadcast reception, and may not give good results in the more demanding circumstances of the LF and MF amateur bands.Several low frequency adaptors have been manufactured for professional HF receivers (eg Racal RA37, RA137).

Fig 10.7 shows a simple homebrew LF/MF to HF converter that has been used successfully at M0BMU, mostly for portable reception in conjunction with a Yaesu FT-817. This uses a 4MHz crystal oscillator and broadband diode mixer module to up-convert input signals from a few kilohertz up to the 550kHz cut-off frequency of the input low-pass filter, to an output range of 4.00 - 4.55MHz. The LF/MF input signal is fed into the DC-coupled IF port of the SBL-1 mixer, and the HF output is taken from the RF port - this allows input frequencies below the 500kHz minimum of the RF port.

The crystal oscillator uses one gate of a 74HCU04 hex CMOS inverter IC, with the remaining five gates used as a buffer amplifier to drive the diode mixer. A wide range of other crystal frequencies could also be used to obtain different output frequency ranges if preferred; due to the broad-band nature of the mixer, no further modification is needed other than to change the crystal. An oscillator frequency below 2MHz makes the circuit more susceptible to IF breakthrough and image responses, while frequencies much over 10MHz will lead to reduced frequency stability, which may be a problem when narrow-band modes such as QRSS are being received.

The converter output has a simple -3dB attenuator pad to reduce the effect of output impedance variations on the mixer; the circuit therefore has an overall loss of about 10dB. No post-mixer amplifier stage is included, since most HF receivers include a low-noise preamplifier that can be switched in to perform this function.

The crystal frequency can be adjusted by setting the HF receiver to exactly 4MHz (or other crystal frequency), and adjusting the trimmer capacitor so that the oscillator signal is exactly centred in the CW passband of the receiver. The received input frequency is then the value displayed by the receiver, minus the crystal frequency. This converter gives good performance from 550kHz down to very low frequencies, and for receivers with very poor sensitivity at LF or MF will often give better results than the addition of high-gain preamplifiers. It is also an effective way of extending the low frequency capability of HF-only receivers.

The prototype converters were either built on ground-plane prototyping board, or ‘dead bug style’ on un-etched PCB board. The only setting up required is adjusting the oscillator frequency. The frequency drift of the prototypes was of the order of 1 or 2Hz over a period of hours, which is adequate for all except the most extreme narrow band modes.

This converter has been used with wire antennas and the tuner/preamp of Fig 10.6. For small loop antennas, the preamp circuit of **Fig 10.8** has been used to increase the signal level by about 20dB.

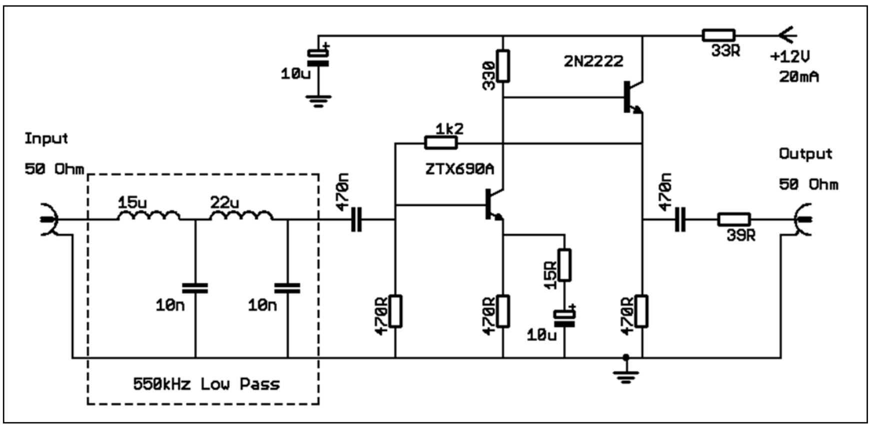


Fig 10.8: 20dB preamplifier

'SOFTROCK LITE' MODIFICATIONS FOR 136kHz AND 500kHz

The SoftRock software-defined radio kits designed by Tony Parks, KB9YIG, have been popular as an entry-level SDR project for the HF bands (see the SDR chapter). The same concept, with suitable modifications, is well suited to the 136kHz and 500kHz bands. Due to the narrow width of these bands, the whole band is easily covered using a PC sound card with the standard 48kHz sampling rate and suitable choice of local oscillator frequency.

The modifications described here were applied to the SoftRock Lite v6.2 kit, but other SDR boards could be modified in a similar way. **Fig 10.9** shows the modifications to the SoftRock circuit (complete documentation of the original circuit is supplied with the kits and can also be downloaded, see [11]). Two major changes are required for 136kHz or 500kHz; the local oscillator frequency is changed, and the input bandpass filter is re-designed.

For the 500kHz band, the crystal frequency is changed to 4.0MHz, and for 136kHz a 1MHz crystal is used. With the 'x8' jumper on the PCB fitted, this gives local oscillator frequencies of 500kHz and 125kHz respectively, giving tuning ranges of 476 - 524kHz, and 101 - 149kHz with a 48kHz sample rate. The existing SoftRock oscillator/driver circuit just works at 4MHz, but not 1MHz. Fig 10.10 shows changes to the circuit to obtain better oscillator waveforms at the lower frequencies. 4MHz crystals are easily obtained; the author had several 1MHz crystals in the junk box, but these are less readily available. An alternative would be to use a 1MHz DIL oscillator module, with its logic-level output connected directly to the divider flip-flop.

The original SoftRock input filter could be re-designed for the LF/MF bands, but the components required would be too big for the tiny PCB, and the rejection of LF and MF broadcast stations near harmonics of the local oscillator frequency would probably not be good enough. Instead, the original tuned input transformer T1 is replaced by a wideband ferrite-cored transformer, and off-board filters were designed with increased rejection at harmonic frequencies. As in the original SoftRock design, component values are fixed, and no adjustment is required. Values are shown for each band. For the 500kHz band, small axial-leaded inductors the size of half watt resistors were used and were quite satisfactory. For the 136kHz band, this type of component has insufficient Q, leading to high insertion loss and a poorly-defined filter passband. Instead, slightly larger radial-leaded chokes wound on ferrite bobbins (eg Panasonic ELC series [12]) were used, which, with Q of around 40, were adequate.

Both 136kHz and 500kHz versions have been used mainly with I2PHD's *Winrad* software [7] and DL4YHF's *Spectrum Lab* software [6], both with excellent results. If the connections to the PC sound card input are made according to the details on the SoftRock kit schematic, the 'reverse I and Q channels' option should be selected to give correct sideband selection. The phase and amplitude balance between I and Q channels should be adjusted to obtain maximum rejection at the image frequency of the centre of the amateur band (about 497.5kHz for the 501-503kHz band, 113.2kHz for 136kHz). This is especially important on 136kHz, where strong signals are present around the image frequency.

Sensitivity on both bands was around 1µv for 10dB SNR with a CW bandwidth of 300Hz; this is not particularly high, but is quite adequate if a transmitting-type antenna is used, or a preamplifier or preselector such as the one in Fig 10.7. In order to prevent low frequency noise generated by the PC and its power supply from getting into the sound card audio input, it was important to maintain isolation between the audio output ground and the RF input ground of the SoftRock circuit. The RF ground was connected to the metal case housing the SoftRock board and input filter, and the audio output and DC supply ground connections were insulated from the case. It was also necessary to use separate DC supplies for the SoftRock and preamplifier or transmitter to prevent ground loops. Connecting RF and Audio grounds together within the case resulted in a 30 - 40dB increase in noise level.

TRANSMITTERS FOR 136kHz & 500kHz

Although many amateur HF transceivers can receive 136kHz and 500kHz signals, they cannot generate useful transmitter power at these frequencies due to PA limitations. There are a few commercially available transmitters for these bands [13, 14], although these have rather limited power output. Therefore, most LF/MF amateur stations use home-made transmitters or transverters.

Due to the very low efficiencies of typical amateur antennas for the 136kHz and 500kHz bands, transmitters are usually

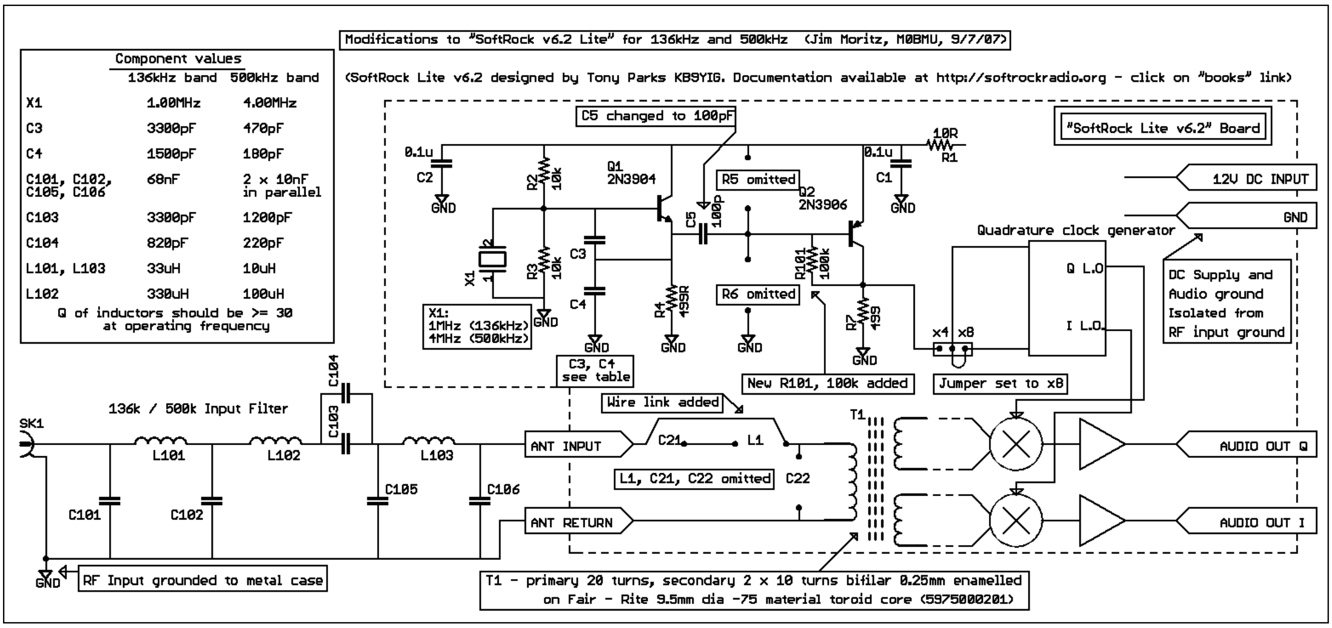


Fig 10.9: Modifications to SoftRock Lite v6.2 for 136kHz and 500kHz

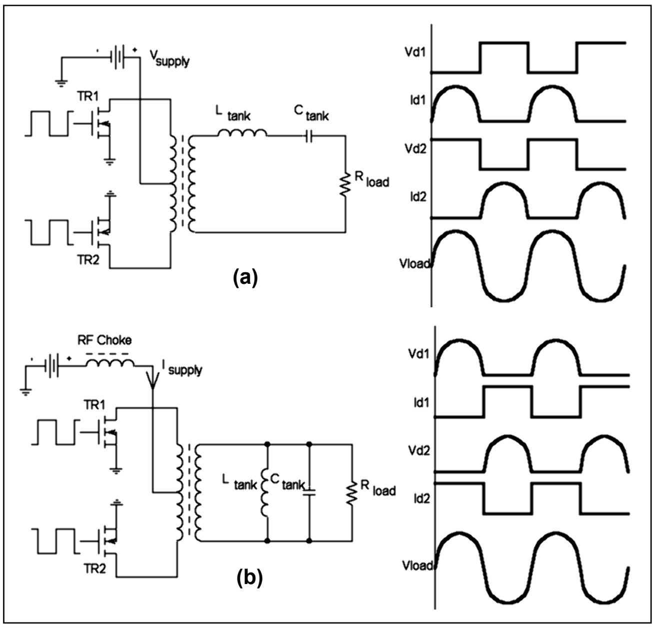


Fig 10.10: (a) Voltage-switching class D amplifier, and (b) Current-switching class D amplifier with MOSFET drain waveforms

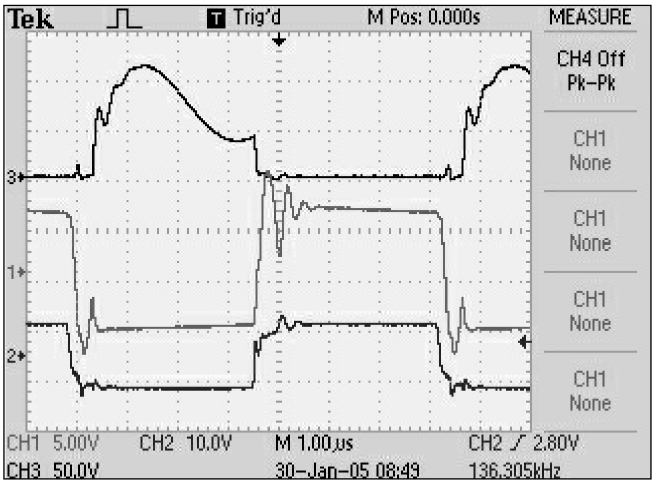


Fig 10.12: Class D waveforms. Upper trace, drain voltage; middle trace, drain current; lower trace, gate drive voltage

required to produce between 100W and a few kilowatts of output. One approach to LF transmitter design is to use HF circuit techniques, with appropriate scaling of components for a lower operating frequency [15, 16]. However, most LF operators are currently using transmitters with switching-mode output stages, operating in class D or class E modes. These can achieve high output powers using quite simple circuits, along with very good efficiency, which considerably simplifies cooling problems associated with high-power linear amplifiers. These circuits are also well suited to inexpensive power MOSFETs and other components intended for switch-mode power supplies operating in a similar frequency range.

Switching mode circuits are, however, more difficult to key or modulate satisfactorily for so-called 'linear' operating modes. Fortunately, most LF operation uses simple on-off keying of the transmitter, or frequency-shift keying. Class D and E PA stages have seen relatively little use for the higher frequency amateur bands; therefore, a description of their design and operation is given below, together with practical designs for complete transmitters. Further LF/MF transmitter designs can be found at [17, 18].

Class D Transmitters

Class D amplifiers (sometimes referred to as 'tuned class D' to differentiate from class D audio amplifiers) fall into two distinct types, voltage-switching, Fig 10.10(a), or current-switching, Fig 10.10(b). In each case, the load is connected to the output stage via a resonant tank circuit. The voltage-switching type has a series-tuned tank circuit; the switching MOSFETs develop a square-wave voltage at the input side of the series tank circuit, however the tank circuit ensures the current flowing in the load is almost a pure sine wave.

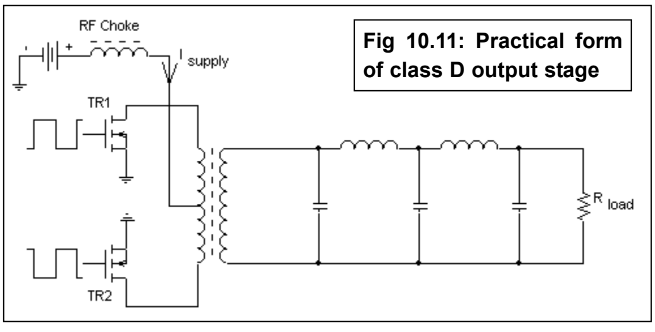


Fig 10.11: Practical form of class D output stage

The current-switching type has a parallel-tuned tank circuit; the supply to the output devices is a constant current which is applied to the tank circuit in alternate directions depending on which MOSFET is switched on. The resulting square wave current applied to the tank circuit again results in an almost sinusoidal voltage across the load. Since a constant-current DC supply is not very practical, a constant voltage supply is used with a series RF choke. Provided the impedance of the choke is much greater than the load resistance, the supply current is almost constant. The major advantage of class D is that the MOSFETs are either fully 'on', in which case the only power loss is due to the MOSFET 'on' resistance, $r_{DS(on)}$, or fully off, with essentially zero power dissipation. In practice, there are additional losses, but these are small compared to linear amplifiers, and efficiency can exceed 90%.

In many amateur circuits, the tank circuit is replaced by a low-Q low-pass filter, Fig 10.11. This circuit is 'quasi-parallel resonant'; it provides a resistive load at the output frequency, but a low shunt impedance at the harmonics. The low Q leads to non-ideal class D operation in that the voltage waveform is not a perfect sine wave, but has the advantages that smaller inductors and capacitors are required, tolerances are less critical, and better rejection of higher harmonics is provided by the multiple filter sections. The voltage and current waveforms of a real-world class D output stage using this circuit are shown in Fig 10.12 Compared to the idealised waveforms, some high frequency 'ringing' is visible. This is due to stray capacitance and inductance which inevitably exists in the circuit, and is undesirable since it causes increased losses, as well as the potential for generating high-order harmonics. It is therefore important to minimise stray reactance, two important causes of which are the parasitic capacitance of the MOSFETs themselves, and the leakage inductance of the output transformer. Adding damping RC 'snubber' networks can also usefully reduce the level of ringing.

As with other types of amplifier, the output power of class D amplifiers is defined by the supply voltage V_{cc}, output transformer turns ratio n and load impedance R_L. For the voltage switching amplifier:

$$P_L = \frac{8n^2V_{cc}^2}{\pi^2 R_L}$$

While for the current-switching class D:

$$P_L = \frac{\pi^2 n^2 V_{cc}^2}{8R_L}$$

These formulas assume that losses in the circuit are negligible; in practice, some losses do occur but since they are small, the results given by the formulas are reasonably accurate.

Class D PA Design Example

The design process for a class D transmitter output stage is best illustrated by an example. The following design is for a LF transmitter with about 200W output, using a current-switching class D circuit. This is a modest power level for 136kHz, but the principles discussed have been applied equally well to designs with 1kW or more output using this circuit configuration, which is probably the most popular in use at present. The complete circuit is shown in **Fig 10.13**.

The first design decision is what DC supply voltage to use, since the power supply is normally the most expensive and bulky part. In this case 13.8V was selected; it can use the standard DC supply found in many amateur shacks. The DC input power required will be about 10% greater than the RF output due to losses, so the expected supply current will be 220W 13.8V = 16A, a level that most 13.8V supplies can readily deliver. For higher power designs, 40 - 60V is often a good compromise, since the problem of large DC and RF currents is then reduced. It is perfectly possible to use an 'off line' directly rectified AC mains supply with no bulky mains transformer, as has been done by G4JNT [19]; note that design for electrical safety is absolutely critical in this case. Inexpensive switching MOSFETs are available suitable for any of these supply voltages. In the ideal push-pull current switching circuit, the peak MOSFET drain-source voltage will theoretically be π times the DC supply voltage, in practice about four times the 13.8VDC supply is likely. MOSFETs should be selected so that only a few percent of the DC input power will be dissipated in their 'on' resistance, $r_{DS(on)}$. This condition also ensures the MOSFET will have adequate drain current rating.

STW60NE10 devices were used for TR2, TR3, with BV_{DS} of 100V, and a typical $r_{DS(on)}$ of 0.016 ohms, leading to about 4W dissipation due to the 'on' resistance and 16A supply current (I^2 x $r_{DS(on)}$). Additional dissipation occurs during the transient period where the device is switching 'on' or 'off'. This can be determined from measurement of circuit waveforms, but can be assumed to be similar to that due to $r_{DS(on)}$. In normal operation therefore, each MOSFET will only dissipate a few watts; however with a severe mismatch, power dissipation can be much higher, especially without DC supply current limiting. For a robust design, the MOSFETs and their heatsink should be able to dissipate of the order of 50% of the total DC input power, at least during a short overload period. The STW60NE10 devices have a TO-247 package and can dissipate 90W each at a case temperature of 100 degrees C, which is adequate.

The output transformer is the most important part of the design. It is normally wound on a core using the same ferrite grades that are used for switch-mode power supplies. These may be large toroidal cores, pot cores, or 'E' cores with plastic bobbins. Several manufacturers produce suitable materials; these include Ferroxcube (Philips) 3C8, 3C85, 3C90, Siemens N27, N87, Neosid F44, and Fair-Rite #77 grades. All these ferrites have permeability around 2000, and have reasonably low loss at 137kHz. They are available in a variety of forms, such as EE, EC, and ETD styles, and sizes; a designation such as ETD49 means an ETD style core that is 49mm wide. A good selection of different core types is available from component distributors [12, 20] at reasonably low cost.

Transformer design is a complex topic in its own right, but a simplified procedure usually gives satisfactory results for amateur purposes, as follows. Given the supply voltage and load impedance (usually 50 ohms), the turns ratio of the output transformer determines the output power. Rearranging the formula for current-switching class D given in the previous section gives:

$$n = \sqrt{\frac{8P_L R_L}{\pi^2 V_{cc}^2}}$$

For V_{cc} = 13.8V, R_L = 50Ω, and P_L = 220W, this gives n = 1:6.8.

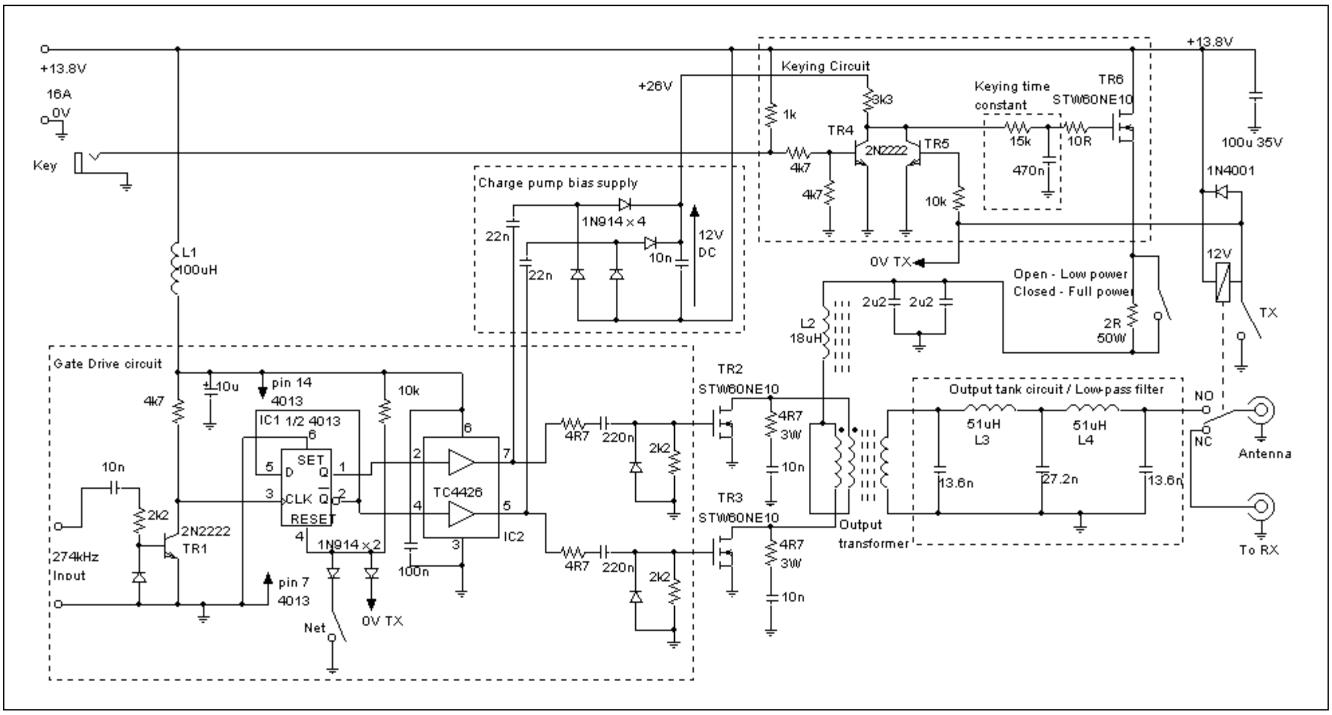


Fig 10.13: 200W class D transmitter circuit

Next, a suitable sized core is chosen. As a guide, using the types of ferrite listed above, an ETD34 core is suitable for powers up to 250W, an ETD44 core for 500W, and an ETD49 for up to 1kW. Similar sizes in different styles have similar power handling. If in doubt, use a bigger core!

The number of turns N in the secondary winding can then be determined. N must be large enough to keep the peak magnetic flux B_{peak} to a value well below the saturation level at the expected output voltage level, V_{RMS}:

$$B_{peak} = \frac{V_{RMS}}{4.44 fNA_e}, \quad V_{RMS} = \sqrt{P_L R_L}$$

Where Ae is the effective area of the core in m^2. The number of turns must also be large enough so that the inductance of the winding has a large reactance compared to the load impedance. A value of X_L about 5 - 10 times the load impedance is desirable. An ETD34 core and bobbin of 3C85 ferrite material was available, which according to the manufacturer's data has A_e of 97.1mm^2 (97.1 x 10-6 m^2), and A_L of 2500nH/T^2. A suitable maximum value of B_{peak} for power-grade ferrite materials is around 0.15 tesla. For 220W output, V_{RMS} is 105V. A few trials using the formulas resulted in n = 14 turns, B_{peak} = 0.127T and L = 490uH, X_L = 422Ω, which meets the criteria given.

Two turn primary windings result in a turns ratio of 1:7, close enough in practice to the 1:6.8 design value. The primary windings were 4 x 2 turns, quadrifilar wound of 1mm enamelled copper wire, using two windings in parallel for each half of the primary winding. The secondary of 14 turns of 0.8mm enamelled copper was wound on top of the primaries, and insulated from them with polyester tape.

The DC feed choke L2 must be capable of handling the full DC supply current without saturation and also have a high reactance at 137kHz compared to the load impedance at the transformer primary, which is (50Ω / 7^2), about 1Ω. A reactance of 10Ω or greater is adequate, requiring at least 12uH. A high Q is not required, since only a small RF current flows in the choke. The 18uH choke used a Micrometals T-106-26 iron dust core. Iron dust cores of similar types to this can often be salvaged from defunct PC switch-mode PSUs. The winding used 2 x 17 turns in parallel of 1mm^2 enamelled copper wire. An air-cored inductor would also be feasible, if more bulky.

The output filter consists of two identical cascaded pi-sections. The filter should provide a resistive load at the 137kHz output frequency, but a low capacitive reactance at harmonics. This can be achieved by designing the pi-sections as low-Q matching networks, with equal source and load resistances. This yields a circuit with two equal capacitors. The standard pi-section design formulae can be used, modified for $R_{in} = R_{out}$ = R:

$$X_C = \frac{R}{Q}, \quad X_L = \frac{2QR}{Q^2+1}$$

$$C = \frac{1}{2\pi f X_C}, \quad L = \frac{X_L}{2\pi f}$$

Most designers select Q between 0.5 and 1. Metallized polypropylene capacitors are a good choice, since they have low losses at 137kHz, and are available with large values and high voltage ratings. The DC voltage rating should be several times larger than the RMS RF voltage present; the main limitation is the heating effect of the RF current causing internal heating of the capacitor. Several 6.8nF, 1kV polypropylene capacitors were available, so C = 2 x 6.8nF = 13.6nF was used. This has reactance of 85.4Ω at 137kHz, forcing Q = 0.585, and giving X_L = 43.6Ω, L = 50.6uH. The inductors must have low loss at 137kHz to avoid excessive heating. Micrometals T130-2 iron dust cores were used, wound with 68 turns of 0.7mm enamelled wire. It is a good idea to check the capacitance and inductance of the filter components using an LCR meter or bridge. However, the main effect of small errors is only to slightly alter the output power from the circuit, without greatly affecting the efficiency.

The drive signal applied to the class D output stage is a 50% duty cycle square wave. The 137kHz gate drive signal is obtained from a 274kHz input using a D-type flip-flop in a divide-by-2 configuration, guaranteeing an accurate 50% duty cycle. When the circuit is switched to receive, the flip-flop is disabled by pulling the reset input high, preventing a 137kHz signal leaking to the receiver input and causing interference. For netting the transmitter, the 'net' switch enables the flip-flop. The MOSFETs require zero or negative gate voltage to switch the transistor fully off, and +10 to +15 volts to bias them fully on. The MOSFET gates behave essentially as capacitors, requiring transient charging and discharging currents as the drive voltage switches on and off, but drawing no current while the gate voltage remains stable. In order to achieve fast MOSFET switching, a TC4426 gate driver IC is used. These driver ICs accept a TTL-compatible logic level input signal, and are designed to produce peak output currents of 1A or more which charge and discharge the MOSFET gate in a fraction of a microsecond. A disadvantage of using a flip-flop to generate the drive signal is that ,if the input signal is lost, one MOSFET will remain switched on and act as a virtual short across the supply. To avoid this, the gate driver is capacitively coupled to the MOSFETs; the shunt diodes perform a DC restoration function, making the full positive peak voltage available to drive the gate. If drive is lost, the gates discharge through the 2.2k resistors, switching both MOSFETs off. The 4.7 ohm resistors in series with the gate drivers help to reduce ringing.

Each MOSFET has a series RC damping network from drain to source, reducing high frequency 'ringing' superimposed on the drain waveform. The component values are best determined experimentally, since they depend on the individual circuit. A good starting point is to make the capacitor about five times larger than the MOSFET output capacitance. A resistance between 2 and 20 ohms is usually effective. Effectiveness is best checked by examining the MOSFET drain waveforms with an oscilloscope, and compromising between minimising high-frequency ringing and excessive power dissipation in the resistors. Larger capacitors and smaller resistors normally result in reduced ringing, but increased dissipation. These components should be appropriate for high frequency use; in this circuit, 4.7 ohm, 3W metal film resistors, and 10nF, 250V polypropylene capacitors were satisfactory.

The transmitter is keyed using series MOSFET TR6. The MOSFET should have low $r_{DS(on)}$ to minimise loss when switched on. A third STW60NE10 was used, although since the maximum voltage applied to this device is the 13.8V DC supply, a lower voltage device could be used instead. During the rise and fall of the keying waveform, dissipation in the keying MOSFET peaks at about 25% of the maximum DC input, 55W in this case. However, when the MOSFET is fully on, it dissipates only a few watts due to $r_{DS(on)}$, and when fully off dissipation is practically zero, so the *average* power dissipated is small, under 10W in this circuit, provided it is not keyed very rapidly. In order to bias this MOSFET fully on, a voltage around 10V higher than the 13.8V DC rail must be applied to the gate.

Only a few milliamps bias are required; a small auxiliary DC supply could be used in a mains-powered transmitter, but in this case the bias voltage was obtained by rectifying the capacitively-coupled gate drive waveform using a charge-pump circuit. The bias is controlled by the key input via transistor TR4, and the keying waveform is shaped by the RC time constant to give around

10ms rise and fall times. TR5 maintains the keying 'off' when the circuit is switched to receive. The 15V zener across the MOSFET gate and source limits the gate voltage to prevent damage. The output power of a class D transmitter can be varied either by changing the supply voltage, or by having multiple taps on the output winding; both these techniques are used in the G0MRF and G3YXM designs described later.

In this design, a resistor can be switched in series with the DC supply, reducing the supply voltage to the output stage to around 4V, and RF output to 18W, for tuning-up purposes. The 2 ohm wirewound resistor dissipates nearly 40W in low power mode, so greatly reduces efficiency, but does make it very difficult to damage the PA due to its inherent current limiting, useful when using a battery supply or when initially testing the circuit.

Construction of this type of transmitter is reasonably non-critical. The low-power parts of the transmitter can be assembled using Veroboard or similar, but the gate driver IC must have a 0.1µF ceramic decoupling capacitor directly across the supply pins due to the large transient currents present. Also for this reason, the gate leads, and the ground return from the MOSFET sources should be kept very short (<30mm). The MOSFETs, output transformer, keying circuit and DC feed carry heavy currents, so connections should be as short as possible and use thick wire (at least 2.5mm^2). The RC damping components should be mounted directly across the MOSFET drain and source pins, and the connections to the output transformer kept short. The circuit described above was assembled on an aluminium plate about 160 x 200 x 3mm, which provided ample heatsinking for the three MOSFETs when air was allowed to circulate freely. In an enclosed box, a fan would probably be desirable.

Testing a class D circuit should start by checking operation of the gate-drive circuit, ensuring that complementary 12Vp-p square waves are present at the MOSFET gates at the correct frequency. A dummy load is almost obligatory for testing LF transmitters (see section on LF measurements). If possible, apply a reduced DC supply voltage to the output stage (but *not* to the gate drive circuit!), or use a series resistor to reduce the supply voltage, as included in this circuit. An oscilloscope is the ideal tool to check the correct waveforms are present. A useful check is the efficiency; the ratio of RF output power to DC input power should be well over 80% if the circuit is working correctly.

Class E Transmitters

Class E power amplifiers are another form of switch-mode output stage that have been successfully applied to 136kHz and 500kHz amateur transmitter construction. The circuit was invented by Nathan Sokal, WA1HQC [21]. The basic class E circuit is single-ended, as shown in **Fig 10.14**. The single switching MOSFET drives a tank circuit C1, L2, C2 with series and shunt capacitors, and is fed with DC supply current via a high impedance choke L1.

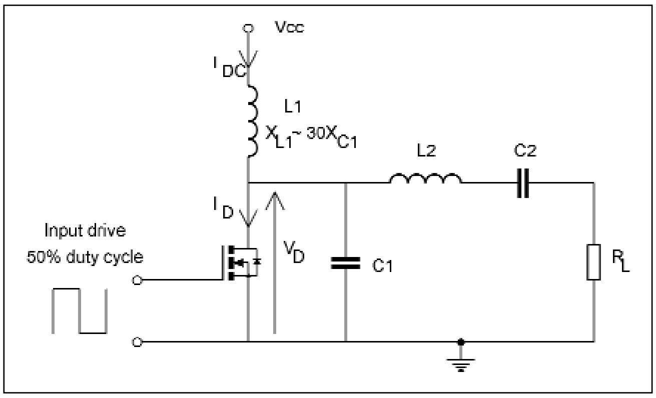


Fig 10.14: Basic class E PA circuit

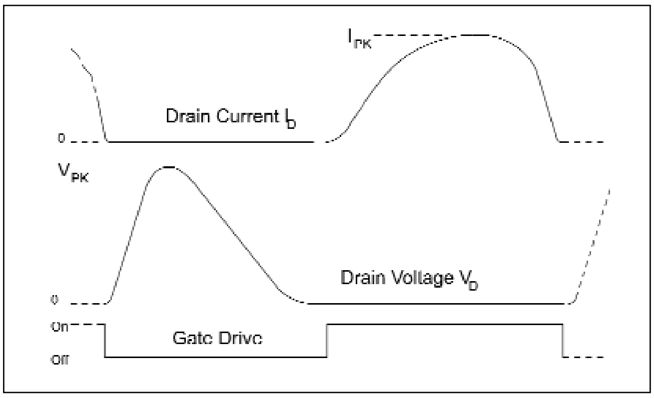


Fig 10.15: Class E waveforms

As in class D PAs, the active device (typically a power MOSFET) is used as a switch that is either fully 'off' or 'on' and thus, in an idealised circuit, no power is dissipated. In switch mode circuits in general, some small losses will occur, partly due to the unavoidable 'on' resistance of the MOSFET, but also partly due to the finite time taken for switching to occur. During switch on, the current is rising while the voltage falls and the reverse occurs during switch off. The instantaneous power dissipation can be high during these short transition periods, since both voltage and current in the MOSFET is simultaneously quite large. The class E tank circuit design aims to minimise these losses by shaping the voltage and current waveforms in the MOSFET so that:

- The MOSFET current decreases to zero before switch-off occurs, so the MOSFET current is zero while the voltage rises
- The voltage across the MOSFET reaches zero before switch-on occurs, so current in the MOSFET does not rise until the voltage is zero

The approximate waveforms of the voltage and current are shown in **Fig 10.15**. The output waveform delivered to RL is effectively filtered by the series-tuned circuit formed by L2, C2 and is an almost pure sine wave. Analysis of this circuit is complex due to its non-linear nature and the non-sinusoidal waveforms. A good description, together with detailed design formulas has been provided in [21]. An approximate design procedure is as follows. Initially, the operating frequency f, supply voltage V_{CC}, output power P_L and tank circuit loaded Q, Q_L are selected. Typically Q_L is between 5 and 10. Referring to Fig 10.15, the component values are approximately:

$$R_L \approx 0.577\frac{V_{CC}^2}{P_L};\ C_1 \approx 0.184\frac{1}{2\pi f R_L}$$

$$C_2 \approx \frac{1}{2\pi f Q_L R_L};\ L_2 \approx \frac{Q_L R_L}{2\pi f}$$

The DC current supplied to the PA, and the peak values of the voltage and current waveforms of Fig 10.16 are approximately given by:

$$I_{DC} \sim \frac{P_L}{V_{CC}};\ V_{PK} \sim 3.56 V_{CC};\ I_{PK} \sim 2.86 I_{DC}$$

These approximations neglect the effect of losses, finite tank circuit Q_L, the impedance of the feed choke L1, but yield values within typically 20%. To obtain more accurate values, the more complex formulas in [21] should be used; design spreadsheets are available from [22] and [23] to simplify this process. In any case, due to component tolerances and parasitic capacitance and inductance in the circuit, it is generally necessary to adjust the tank circuit components experimentally to their final values while observing the voltage waveform with an oscilloscope. A detailed description of this process is also given in [21]. The value of R_L will generally not be 50 ohms; a ferrite-cored matching transformer may be used to obtain the desired load impedance as for the class D circuit, or the tank circuit may be extended to include an additional L-network as described in the G4JNT design in [23, 24] and below. Additional low-pass filtering for harmonic suppression may also be required.

Compared to class D PAs, perhaps the main attractions of class E designs for amateur LF/MF use include a simpler circuit with single-ended input drive and output, and 'cleaner' waveforms with less possibility of high frequency ringing and switching transients, especially compared to the push-pull current-switching class D circuit. This is probably mostly due to the elimination of the push-pull transformer with its inevitable parasitic inductance and capacitance. At higher frequencies, switching losses in class E PAs are significantly reduced compared to class D, but at 136kHz and 500kHz, efficiencies of well over 80% are possible with both types. The main disadvantages of class E are the need for relatively bulky high Q tank circuit inductors and capacitors, and the attendant requirement for a more precise tuning up procedure.

G4JNT High Power Class E Amplifier

The goal of this project was a low cost, high power class E switch mode amplifier for 500kHz. A power output of 500W running from a 50V supply was decided upon as an aiming point. References [21] and [22] provided all the design details, and a version of the design spreadsheet in [22] was developed that added a series L / shunt C output matching network to raise R_{load} to 50 ohms. The series L is absorbed into either the tank circuit L or C values.

Several IRFP462 FETs were available, rated at 500V, 170W dissipation and $R_{ds(on)}$ of 0.4 ohms. The $R_{ds(on)}$ value is a bit high for the projected power level, so two devices were used in parallel.

The air-cored tank coil was wound with 2.5mm litz wire. Ceramic capacitors proved to have high losses when used in the tank circuit, and the final breadboard uses 3.3nF, 1700V metallised polypropylene capacitors built up in parallel combinations to give the required values and share the RF current.

The completed circuit in **Fig 10.16** has input circuitry consisting of a bandpass filter followed by a line receiver to get a close-to-50% duty cycle square wave from a low level input. For driving the FETs, ICL7667 gate drivers were already available. Since the ICL7667 contains two identical drivers, one was used for each parallel FET to reduce the loading due to the FET's 2000pF of input capacitance amplified by Miller feedback. The 100 microhenry DC feed choke is a large toroid, and must be capable of passing the 10A DC feed current without excessive heating.

The unit was powered initially with a 12V supply to the PA to allow tuning safely without over-stressing PA components. The PA was tuned by adding or removing 3.3nF capacitors from the parallel combinations making up the tank circuit capacitors, while viewing the switching waveform and power output (reference [21] contains a description of the tuning procedure and expected waveforms).

The quantisation of the capacitance values in 3.3nF steps is a bit coarse, but combinations could be found giving close to the right waveform shape (**Fig 10.17**) and output power. With 50V, 10.1A DC input, RF output of 400W was achieved, with the heatsink running cool at this level. The peak drain voltage of 190V is also well within the FET's rating so it may be possible to safely increase the DC supply voltage to achieve the 500W target.

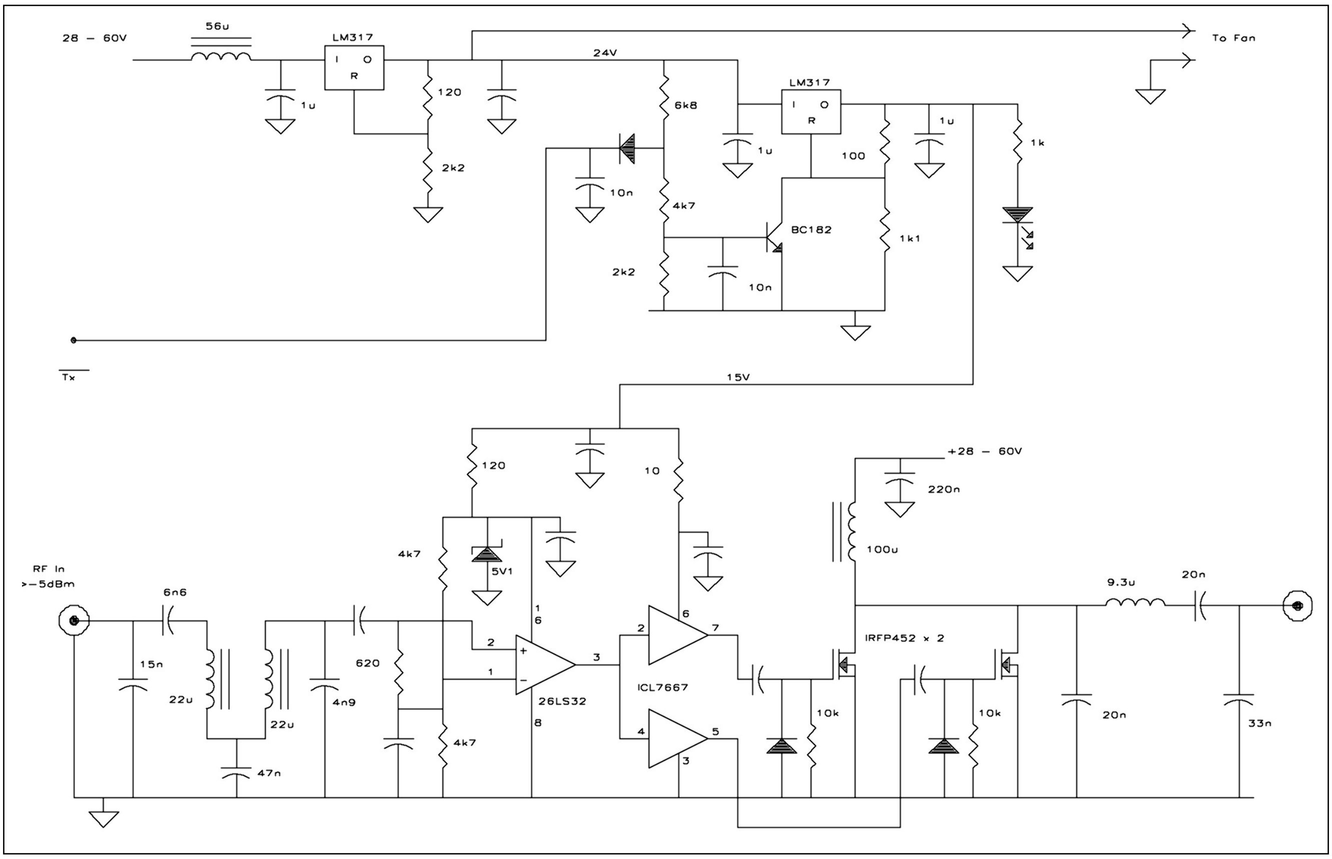


Fig 10.17: G4JNT QRO class E 500kHz PA. All unmarked capacitors are 100nF. Diodes are 1N4145

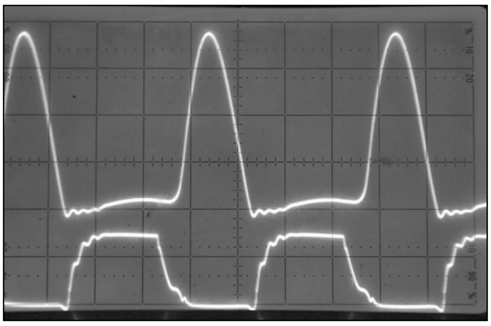


Fig 10.17: Class E PA waveforms: Upper trace - Drain voltage (50V/div); Lower trace - Gate drive voltage (10V/div)

Efficiency at 79% is not the highest that can be obtained from class E PA stages - values in excess of 90% have been claimed. The major power loss in the circuit is due to the relatively high 'on' resistance of the IRFP462 FETs, as can be seen from the approximately 8V V_{ds} voltage drop visible in the waveform of Fig 10.19. More modern devices are available with much lower $R_{ds(on)}$. Also the ICL7667 gate drivers may be marginal in this circuit; again more recent devices are available with higher drive capability. A more detailed article describing this project and the development process can be found at [24], and the tank circuit design spreadsheet incorporating the output matching network can also be downloaded from [23].

IC1	HEF4001
IC2	HEF4017
IC3	HEF4538
IC4	HEF4013
IC5	TC4426
IC6	HEF4023
IC7	7812
Q1, 2, 3, 4	IRFP450
Q5	IRFP260
D1, D2	1N4936
D3, D4, D4, D6	1N4006
Hall effect device	OHN3040U (Farnell 405-656)
BR1, BR2	35A, 600V (Farnell 234-151)
R (for Zobel network "Z")	22Ω, 25W (Farnell 345-090)
Mains transformer	2 x 35V, 530W
T1	Primary 2 x 8 turns, secondary 20 turns, tapped at 12 and 16 turns
CH1	20 turns on 50mm length of antenna-type ferrite rod
T2, T3	Toko 719VXA-A017AO (Bonex)
T4	Primary 1 turn, secondary 2 x 18 turns bifilar
Output filter inductors	54μH. 65 turns 1mm enamelled wire on Micrometals T200-2 toroid core

Table 10.1: Component details for the G3YXM 1kW transmitter

G3YXM 136kHz 1kW Transmitter

G3YXM set out to design a transmitter that is reasonably small, produces around 1kW RF output, and will withstand antenna mis-match and other mishaps. The description here is an abridged version of the original article available via G3YXM's web pages [25]. The major components are listed in **Table 10.1** and the circuit is shown in **Fig 10.18**.

An input signal at 1.36MHz from the VFO (**Fig 10.19**) (from a crystal oscillator and further divider) is divided, the output at IC4 being a symmetrical square wave, driving the output MOSFETs via gate driver IC6. Each MOSFET shown is actually two devices in parallel. The output transformer ratio is set by switch S2. Higher output is obtained with more turns selected. Across the primary of the transformer the Zobel network marked 'Z' (22 ohms and 4n7 in series) reduces ringing. The output is fed to the antenna via a low-pass filter. The cut-off frequency is quite high at about 220kHz as virtually no second harmonic is produced.

The SWR bridge consists of T4 and associated components. It is a bifilar winding of 2x18 turns which forms the centre-tapped

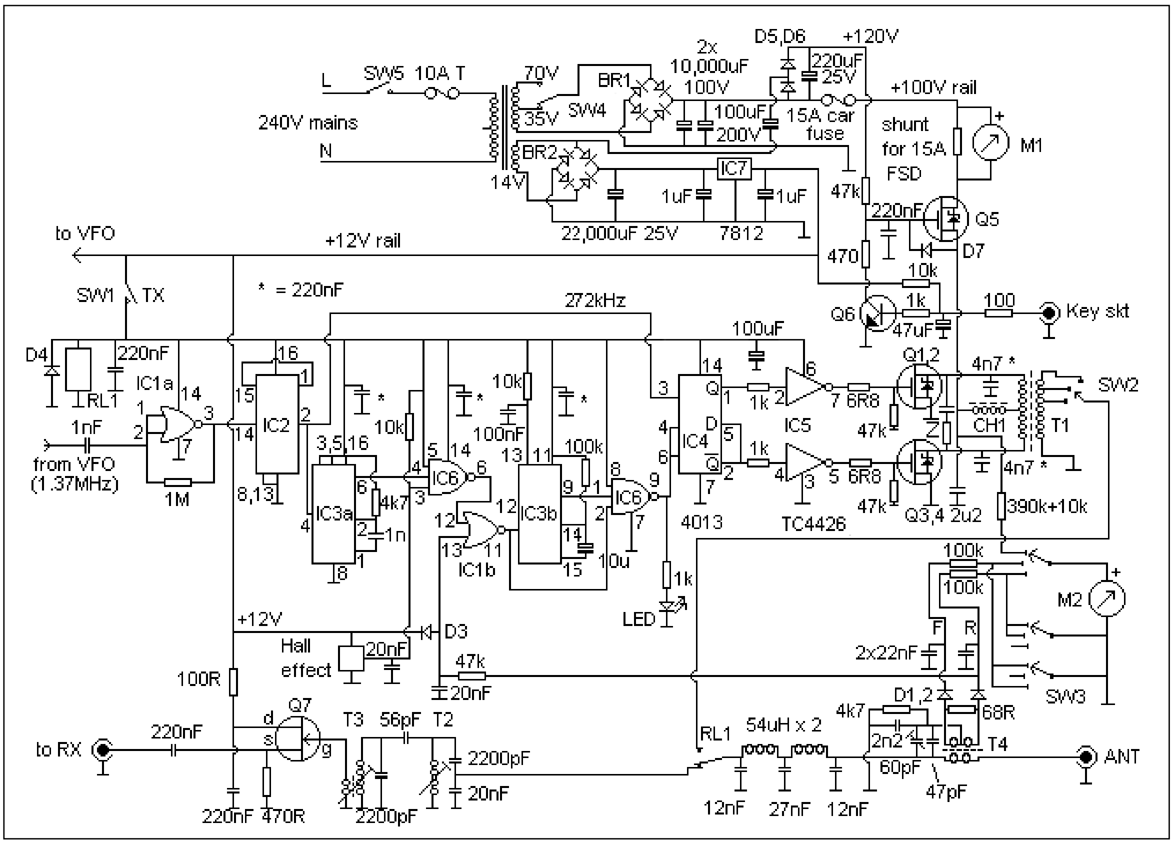


Fig 10.18: G3YXM one kilowatt transmitter

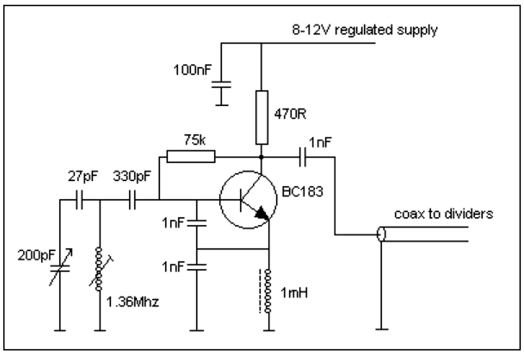


Fig 10.19: 1.36MHz VFO

secondary and the coax inner passing through the toroid core forms the single-turn primary. The protection circuit which cuts the drive for about a second is triggered by high SWR via IC1B, or over-current signal from the Hall-effect device, which is triggered by the magnetic field of CH1, which is made from a 50mm piece of ferrite antenna rod wound with 20 turns of 1.5mm enamelled copper wire. The Hall-effect detector is placed near one end of CH1 and the spacing adjusted to trip at about 20 amps. The receive pre-amp uses coupled tuned circuits giving a band-pass response over the 135 to 138kHz range. A single JFET (Q7) makes up for the filter loss.

The mains transformer used in the power supply has two series 35V windings. The DC voltage is either 50V from the centre tap or 100V from both windings. An auxiliary 12V winding was added by winding 30 turns of 16SWG wire through the toroid. At full output the HT will drop to about 80V. The keying circuit uses a series MOSFET with shaping of rise and fall times to prevent key-clicks. To turn this MOSFET fully on, its gate must be at least 5V positive of its source which is close to the main supply voltage. Diodes D5 and 6 are supplied via a high voltage capacitor from the 12V winding to produce an extra 20V bias for this purpose.

The low-level circuitry was built on strip-board, taking care to keep the tracks short and earth unused inputs. The TC4426 chip IC5 is capable of driving 1.5A into the gate capacitances of the MOSFETs and the decoupling capacitors must be fitted close to the chip with short leads. The 6R8 series resistors are mounted on the gate pins of the MOSFETs, the resistor leads forming the connections to the strip board. It is probably best to use one resistor for each gate. The strip board should be grounded to the earth plane as near as possible to the MOSFETs, which should have the source leads soldered to the ground plane. The two 4n7 capacitors should be connected directly across the MOSFETs. Output transformer T1 should be constructed from two-core 'figure of eight' speaker cable wound eight times through the ferrite toroid, connected as a centre-tapped primary by connecting one end of one winding to the opposite end of the other. The secondary is wound over it with 20 turns of thin wire tapped at 12 and 16 turns. The Zobel network should be wired from drain to drain with short wires.

Get the PSU, VFO and CMOS stages working first. Check with a scope that you have complementary 12V square waves on the gates, the waveform will be slightly rounded off due to the gate capacitance. Connect the transmitter to a 50 ohm load and, having selected the first tap on SW2, apply 50V (SW4 in low position) with a resistor in place of the fuse to limit the current. The MOSFETs should draw no current without 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, either the load is mis-matched, the SWR bridge is connected backwards or the 60pF capacitor needs adjustment. If all seems well, remove the current limiting resistor and increase the power by selecting taps, key the rig in short bursts and check for overheating of MOSFETs and cores. When you are happy that the transmitter 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. The receive preamplifier filter inductors can be aligned using a signal near 137kHz; the tuning is very sharp.

GW3UEP 100W 500kHz CW Transmitter

This 100W MF CW transmitter and the 25W QTX ('Quick-TX', see [26]) were developed by GW3UEP from earlier designs for 160m and 80m. When NoVs for amateur 500kHz operation arrived in 2007,the solution was to hand and the QTX was born!

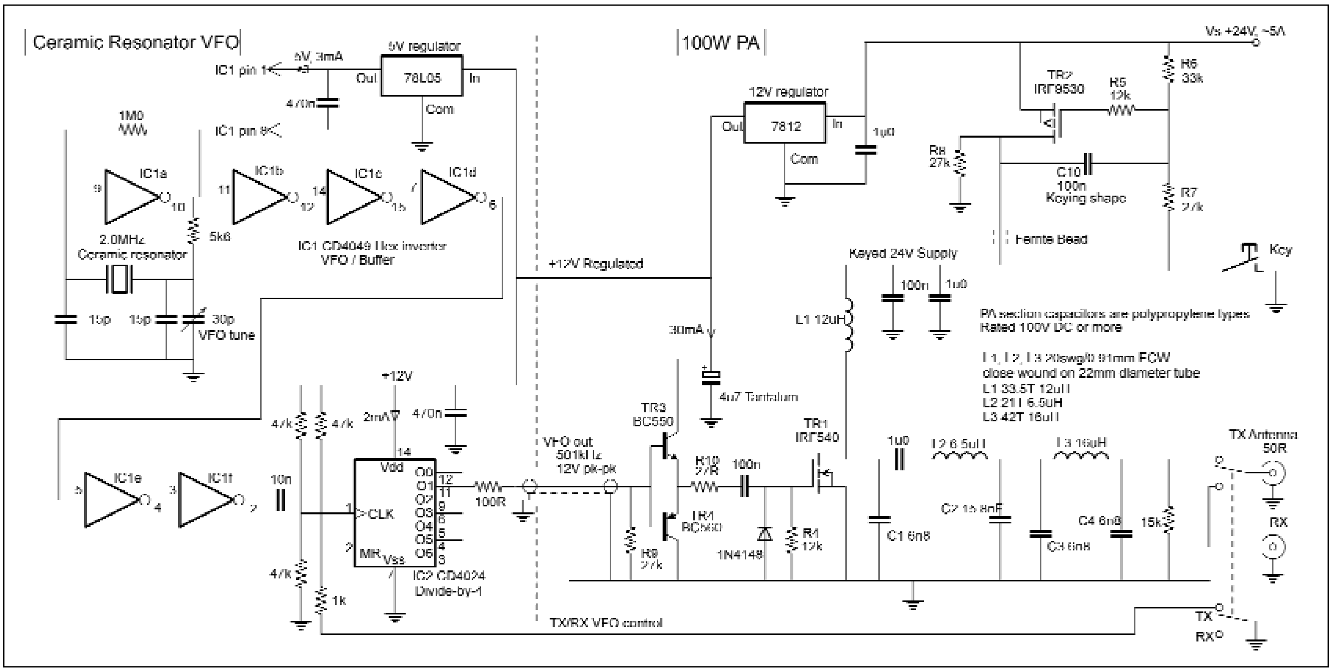


Fig 10.20: GW3UEP 100W 500kHz transmitter

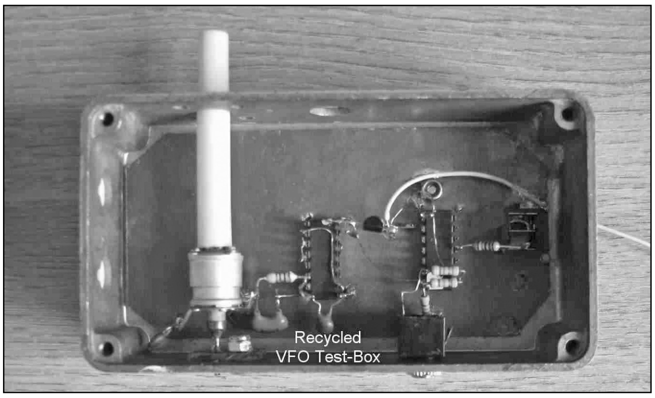


Fig 10.21: Ceramic resonator VFO for GW3UEP transmitter

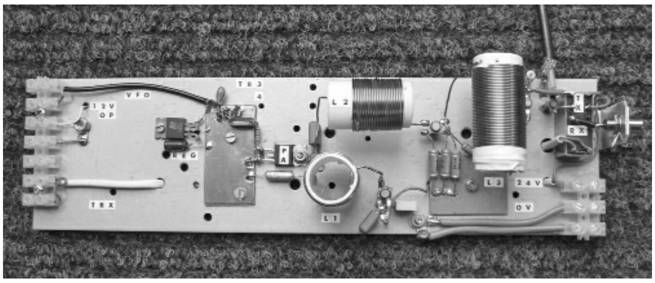


Fig 10.22: GW3UEP transmitter power amplifier

Subsequent increases in permitted ERP to 10W have made greater output power useful, leading to this 100W version. Thanks go to GW4HXO and EI0CF for their collaboration in construction, evaluation and on-air-testing of the transmitters.

The complete schematic for the transmitter is shown in Fig 10.20. A CMOS 4049 hex inverter (IC1) with 2MHz ceramic resonator provides a temperature stable VFO. The buffered 5Vpk -pk output is ac-coupled to a 4024 divider (IC2), which delivers 12Vpk -pk output at 501-504kHz. Frequency stability is optimised by low power / continuous operation of the oscillator circuit. The VFO box (Fig 10.21) is separate from the PA unit (Fig 10.22) in order to avoid thermal coupling and temperature change.

The IRF540 MOSFET was chosen for operation as the 500kHz power switch, TR1. TR3 and TR4 form a zero-biased complementary voltage follower, buffering the IC2 output stage and providing adequate source/sink current for the IRF540 gate charge (BC549, BC559 are also suitable for TR3 and TR4). The gate is AC-coupled to the buffer and DC-restored to ground, to prevent high DC current flow in TR1 should a fault in the 500kHz drive occur. The PA operates in switched-mode with drain efficiency in the 80% range.

The output circuit provides matching and LPF functions, and presents a clean sine wave into the 50 ohm load. C1/L1 forms a resonant MF tank circuit. L-match C2/L2 transforms the 50-ohm output load to a lower impedance at the drain. The output inductors are air-cored and wound on 22mm diameter plastic 'waste pipe' available from plumbing suppliers. PA keying is achieved with P-channel MOSFET TR2, which also shapes the keyed RF envelope and eliminates key-clicks. The key input switches TR2 gate via R6 and R7, which along with C10 also set the rise and fall times. R5 ensures stability by rolling off the frequency response of TR2, forming a LPF with its input capacitance.

The maximum supply voltage for the transmitter is 25V; this allows for a voltage drop across TR2, which has VDS of 1V at 5A supply current. R6/R7 reduce the gate-source voltage of TR2 to 14V with a 25V supply. Heat sinks are required - at maximum output, TR1 dissipates 20 - 25W, TR2 5W.

Ideally, a stabilised 24V PSU with current limiting set to approximately 5A should be used. Additionally, a 5A fuse should be incorporated. An un-regulated PSU should deliver 24-25V maximum on load. Operation over the range 14-24V is recommended. A typical setting-up procedure includes the following steps and approximate values:

- Terminate transmitter output with a 100W dummy load/power meter and observe the DC supply current.
- Remove R10, set to Tx and apply 24V supply: check <20mA total. Check 5V/12V regulator outputs.
- Check 12Vpk-pk at 501-504kHz across R9.
- Check 0V/24V across R8 with key-up/down, <20mA total.
- Restore R10.
- Set supply to 14V.
- Check >10Vpk-pk at 501kHz across R4.
- Key-down: check 30W RF output and 3A supply current.
- Set supply to 24V.
- Key-down: check 100W RF output and 5A supply current.
- Check PA drain waveform is a clean pulse and that efficiency is >80%. The drain voltage waveform should be approximately 100Vpk-pk. (The drain voltage waveform for this circuit is very similar to Fig 10.17, except for the lower voltage swing.)
- Trim L2/L3 for best results. (GW3UEP uses a 'tuning wand' consisting of wooden tooth-picks fitted with a ferrite bead or dust core, and a short circuit loop of wire about 13mm diameter. Inserting the ferrite into the coil increases inductance, inserting the shorted turn reduces inductance.)

This transmitter design is in use by a number of amateur stations; further information and details of design variants can be found at [26,27].

200W Multi-Mode EER Transverter for 136kHz and 500kHz

The two main objectives of this design (illustrated in Fig 10.23) are to provide a convenient way of transmitting and receiving digital mode signals on the 136kHz and 500kHz amateur bands, and to act as a test bed for transmission of different modes using the EER (envelope elimination and restoration) technique for power amplification.

The transverter signal source is an HF SSB transceiver with audio input and output from a PC sound card. This enables the use of a wide range of 'sound card mode' software, together with the convenient VFO and filter facilities of the HF rig. The basic transverter mixes this signal with a 4MHz local oscillator to obtain a low-level output in the range 20kHz - 550kHz. For reception, the signal path is reversed to convert the LF/MF input signal up to the 4.0 - 4.5MHz range. Transmit output at 200W PEP is achieved using a class D PA and modulator using the EER technique, although the transverter also has a low-level output for use with a conventional linear PA.

Envelope elimination and restoration (Kahn technique) is a method of generating amplitude- and phase-modulated signals using a high efficiency, non-linear PA, without introducing excessive distortion. The signal to be amplified is divided into a carrier phase channel and an amplitude envelope channel.

The constant-amplitude, phase-modulated carrier frequency is applied to the PA input, which can be a class D or E switching mode type to achieve high efficiency. The amplitude modulation signal is used to modulate the DC supply voltage to the PA. The output from the PA retains the phase modulation in the carrier signal, but now also varies in amplitude proportional to

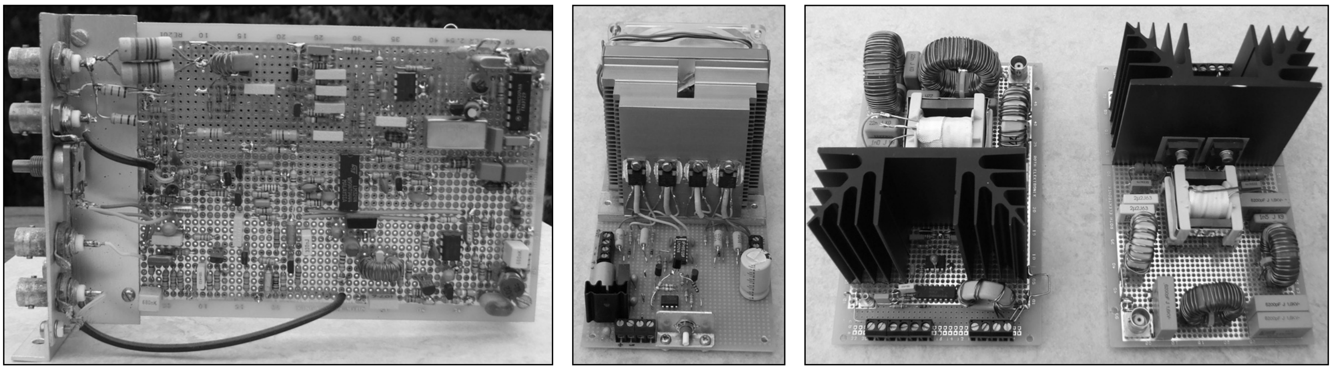


Fig 10.23: Components of the EER transverter system: (left) Low-level transverter; (middle) Series modulator; (right) 136kHz and 500kHz power amplifiers

the envelope modulation signal. Since any type of signal is effectively a carrier frequency with a combination of amplitude and phase modulation, the EER technique can in principle be used with any form of modulation. If the high efficiency amplifier is used in combination with a switch-mode modulator, very high efficiencies are possible, making the technique popular for high-power AM/SSB/digital broadcast transmitters.

For these relatively high power, wide-band transmitters, complicated techniques are required to ensure alignment between amplitude and phase channels. But for LF/MF amateur applications at moderate power levels and using narrow-band modulation, satisfactory signal quality can be achieved quite easily.

This design uses a simple linear modulator which dissipates a significant amount of the DC input power. For types of modulation with a reasonably high crest factor (ie ratio between average power and PEP) the losses in the modulator are quite low. For example the efficiency calculated for some common types of modulation, assuming an idealised system with a 100% efficient PA and a series modulator that delivers the full supply voltage to the PA at modulation peaks gives the results shown in **Table 10.2**.

Mode types	Efficiency
A1A (CW), FSK (no envelope modulation):	100%
BPSK with envelope shaping (eg PSK31):	89%
Two tone ‘IFK’ modulation (eg Throb):	78%

Table 10:2: Theoretical efficiencies achievable using EER

In practice, there are additional losses in both PA and modulator, so in this design actual efficiency is perhaps 20% less than these figures, but this is still better than a class AB linear PA transmitting the same modes. The linear design is simpler than a switch-mode modulator, and power dissipation in the modulator is quite manageable at the 200W output level.

A block diagram of the transverter system is shown in **Fig 10.24**. The driving IC-718 HF transceiver is operated at 4.1 - 4.5MHz IF frequency. This frequency was chosen to utilise some 4MHz crystals that were available; the IC718 can easily be modified to operate on frequencies outside HF amateur allocations.

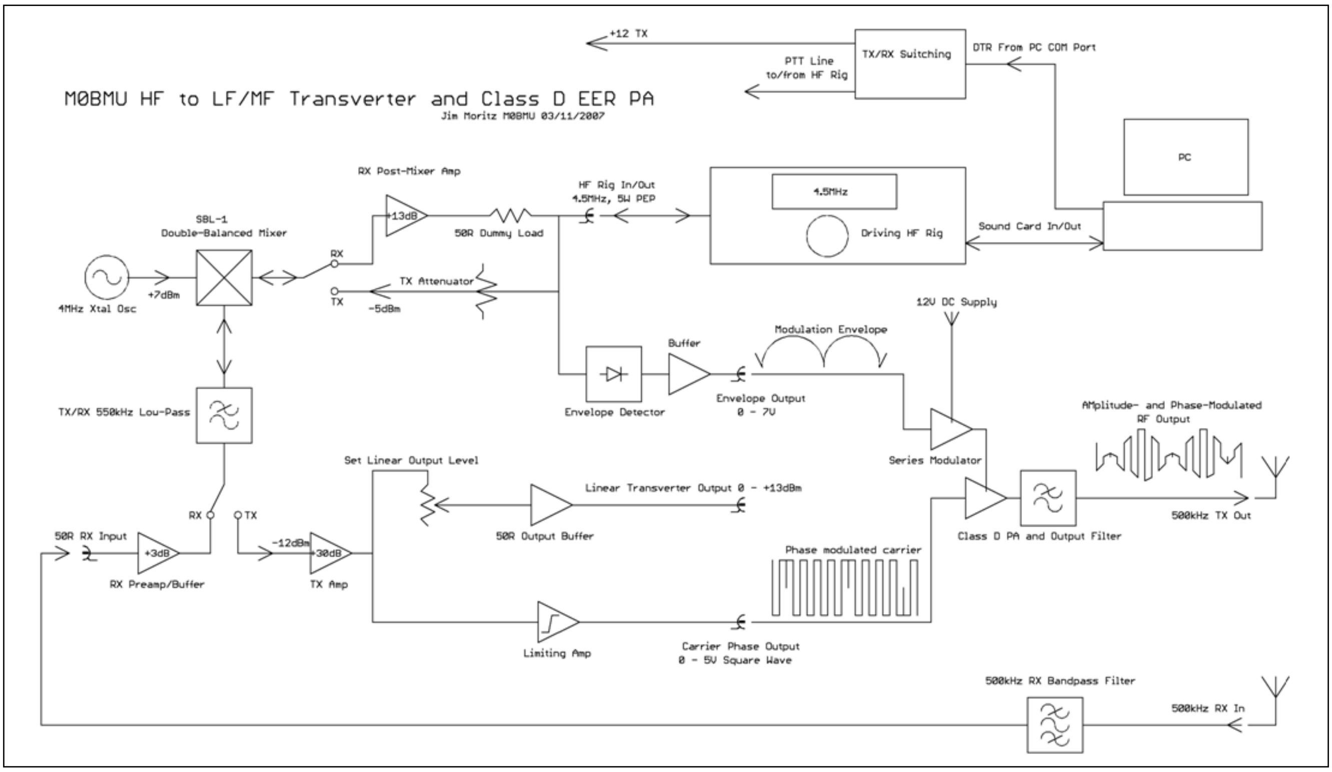


Fig 10.24: EER Transverter block diagram

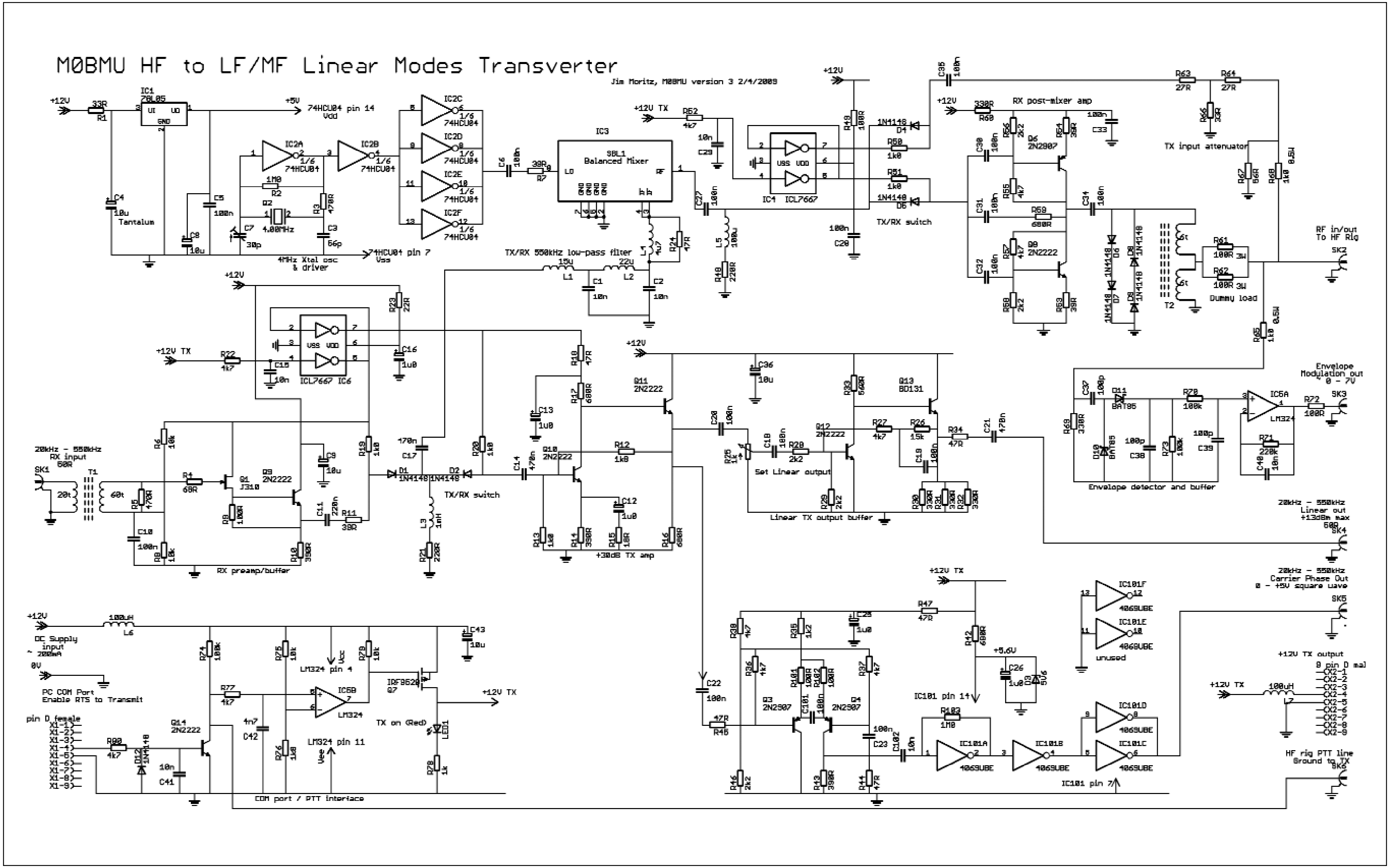


Fig 10.25: Transverter receive and low-level transmit sections

The transverter IF input/output circuit is wide-band, so other input frequency ranges could be used just by changing the crystal frequency. The low-level transverter output is also broad band, and is about 3dB down at 20kHz and 550kHz, allowing it to be used anywhere in this range, although the PAs described are narrow band and limited to frequencies around 136kHz and 500kHz. The receive channel covers a similar frequency range, and achieves about 13dB SNR with a 0.1µV input signal in 250Hz bandwidth. This level of sensitivity allows the use of separate small receiving antennas to combat interference problems.

In the transverter, the carrier phase and amplitude signals are separated from the modulated signal from the HF transceiver. The carrier phase signal is obtained by feeding the down-converted signal into a limiting amplifier. The limiter output is a logic-level square wave at constant amplitude. The envelope modulation signal is obtained by rectifying the HF signal with a diode envelope detector, buffered by an op-amp follower.

The output from the HF transceiver is set to about 5W PEP; this was found to give a clean signal, and the level is high enough to be adjusted easily using the HF rig drive level control. Also, good linearity is obtained from the envelope detector with this high level signal. A buffered linear output at up to +13dBm is also provided.

The modulator and PA are designed for use with 13.8V DC. The objective was to produce a rig for possible future portable battery operation, although it is also convenient for use with the high current 13.8V PSUs present in many shacks. The PA circuits draw about 19A at full output.

Separate PA circuits are used for 136kHz and 500kHz; since the MOSFET and driver components are cheap, and band switching would be quite awkward due to the high currents and low impedances involved, it was felt better to have separate PAs for each band rather than trying to produce a dual band design.

The schematic of the low-level sections of the transverter is shown in **Fig 10.25** The output from the driving HF transceiver is connected permanently to a 50-ohm dummy load, R61, R62.

On transmit, diodes D6 - D9 conduct, effectively grounding the other end of the load resistor via T2. An attenuated portion of the transmitter output is fed to the diode mixer via the diode transmit/receive switch.

On receive, D6 - D9 do not conduct, and the load resistor is in series with the output of the post-mixer amplifier Q6, Q8. This circuit was designed to make the transverter robust against being damaged by accidentally transmitting into the receive circuits, and can tolerate a 5W level indefinitely, and 100W for short periods.

The transmit/receive diode switch uses an ICL7667 driver IC to drive 10mA forward bias, or 2.5V reverse bias into switching diodes D4, D5. IC3, a SBL-1 double-balanced diode mixer is used for down/up conversion. The LF/MF signal enters or leaves via the DC-coupled IF port, to allow operation at low frequencies. The LO signal for the mixer is from a 4MHz crystal oscillator using one gate of CMOS inverter IC2. The remaining gates form a driver for the 50-ohm input impedance of the mixer. The LF/MF mixer input/output is filtered by a 550kHz low-pass filter to remove LO and image components.

The LF signal is switched between transmit and receive paths by another diode switch (D1, D2) - in this case the switch driver also switches the DC bias to the receive preamp and the transmit amplifier. This effectively increases the switch isolation, reducing possible problems with feedback between transmit and receive paths. If an external receive preamplifier was used with this transverter, it would be advisable to arrange that this was also switched off on transmit. The preamp/buffer Q1, Q9 provides a 50 ohm termination at the receive input, and for the mixer and low-pass filter. The gain of this amplifier, and the overall receive gain, is only around 3dB. This minimises problems due to overloading by strong signals, whilst maintaining a reasonably low noise figure.

On transmit, the signal from the mixer is amplified by Q10, Q11. An output is taken from this point via a level-setting pot and buffer Q12, Q13 for optional use with a conventional linear PA. The signal is also applied to a limiting amplifier to generate the carrier phase output. The long-tailed pair Q3, Q4 provide a well-defined symmetrical limiting action. This output is applied to a biased CMOS inverter, IC101a, which gives additional gain at low signal levels, and additional inverters bring levels up to a logic-compatible 0V, +5V. The output square wave is symmetrical within a few percent over the likely input signal range, which minimizes unwanted AM to PM conversion. The envelope modulation is extracted from the HF input signal by envelope detector D10, D11 and buffered by op-amp IC5A. The high signal frequency allows small detector time constants, providing adequate filtering without introducing large phase shifts. Schottky diodes are used for their lower forward voltage drop. The modulation envelope output is about +7V at the modulation peaks.

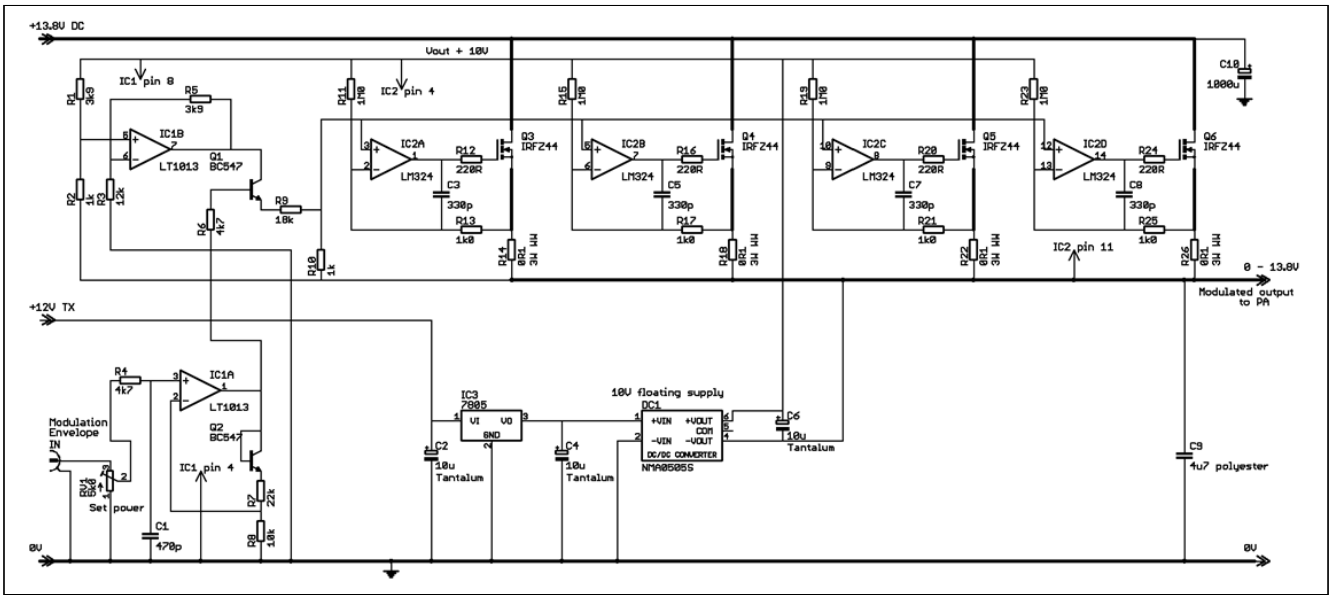


Fig 10.26: 200W series modulator

Transmit/receive switching is controlled via the PC COM port DTR handshake line by most digital modes software. The IC-718 and most other rigs are switched to transmit by pulling their PTT line to ground, which is done by Q14, which also switches MOSFET Q7 on via op-amp IC5B, providing a switched +12V Tx line, which is used for T/R switching in the transverter, PA and elsewhere if needed. This arrangement also operates the +12V Tx line if the PTT line is switched to transmit internally, for instance when operating CW.

The only adjustment required to the transverter circuit is to trim the conversion oscillator frequency. The frequency can be set using an accurate frequency counter; preferably the HF rig frequency should also be trimmed. If a suitable counter is not available, the best approach is to set the HF rig to 4.000MHz CW receive, and trim the transverter oscillator so that the receive audio note is equal to the nominal CW pitch. This can be set within a few Hertz using the spectrogram display in many digital mode programs. This procedure makes any frequency offset in the transverter equal to the offset of the HF rig, minimising the error in the output frequency

The 200W series modulator schematic is shown in **Fig 10.26**,. The modulator has to be capable of dissipating significant power - about 55W peak during normal operation, but reaching nearly 100W during worst-case conditions with the PA output shorted.

Four MOSFETs (Q3 - Q6) in parallel are used. To achieve accurate current sharing, each MOSFET has a current sensing resistor (R14, 18, 22, 26). The op-amps IC2a-d control the gate bias on each MOSFET so that the voltage across the current sensing resistor is equal to the control voltage applied to the non-inverting inputs of the op-amps; thus the current in each MOSFET is the same, and determined by the control voltage between the junction of R9, R10 and the output voltage rail. Since this control voltage is sensed between the terminal of R9 and the output voltage terminal, the overall circuit behaves as a voltage follower, with the output terminal voltage following the voltage applied to R9 with somewhat less than unity gain.

The modulation envelope input is amplified by IC1a and drives R9 via follower Q1. To prevent excessive power dissipation in both modulator and PA under fault conditions, a 'fold-back' current limiting characteristic is desirable, where the maximum output current is reduced at low output voltages, reducing the maximum dissipation that can occur in the modulator. Current limiting occurs because the maximum control voltage is equal to the collector voltage of Q1. This voltage therefore determines the maximum voltage across the current sensing resistors and the maximum output current. The collector voltage of Q1 is generated by op-amp IC1B summing a fraction of the output voltage with a DC offset produced by R1, R2. The voltage varies from about 2.6V with zero output to about 6.8V at maximum output voltage. This sets the current limit; with the given values of R9, R10 and the 0.1 current-sensing resistors to about 8A at zero output voltage, up to about 23A at maximum output. Thick lines in Fig 10.29 represent high-current paths - up to 20A at 200W out.

A small 5V to +5/-5V DC - DC converter provides a 10V floating supply for the MOSFET driving op-amps. The DC-DC converter is supplied from the +12V TX line via IC3, so that it is switched off on receive to prevent possible QRM - the internal switching frequency is about 100kHz.

IC3 has a small clip-on heatsink. The current-limiting op-amps IC2 are only required to operate with inputs and outputs positive with respect to the modulator output, and the LM324 is a 'single supply' type, so the negative supply of this IC is connected to the output rail. IC1 inputs must operate down to zero, so its negative supply pin is connected to 0V, and it must also be a single-supply type. The total supply voltage of this IC is thus about 23.8V maximum, depending on the modulator output. In order to allow possible future use of the circuit with a higher supply voltage such as 24V DC, the LT1013 dual op-amp was chosen, as it allows up to 44V total supply voltage.

The heatsink used for the modulator depends on the type of operation required. For normal transmit/receive use, average power dissipation is low and a 1°C/W heatsink provides adequate cooling, but if long-duration beacon transmissions are envisaged, with possible long periods of overload resulting, a bigger heatsink will be needed. A fairly cheap and compact solution to this requirement is a large 'CPU cooler' heatsink with fan as available from PC parts suppliers. The one used for the prototype (Akasa AK862) was more than adequate, with only about 20°C temperature rise after a prolonged short circuit. Note that the fan has to be running for this type of heatsink to be effective. If a smaller heatsink is used, it would be advisable to attach a normally-closed bimetallic thermal switch with a trip temperature of around 70°C, to disconnect the +12V Tx line if the heatsink temperature is excessive.

It was found that the Eltec SMPS 60/20 power supply used with the prototype modulator showed large fluctuations in output voltage when the modulator was keyed on and off at maximum output, apparently due to poor transient regulation - a large (68000μF) electrolytic capacitor across the PSU output

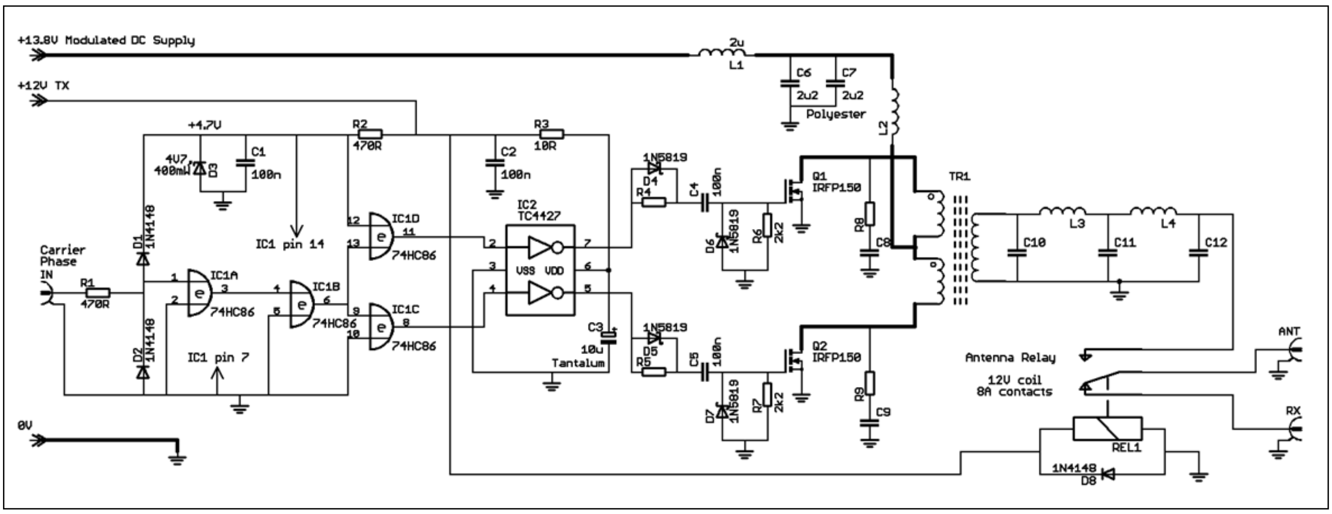


Fig 10.27: 200W class D PA

Q1, Q2	Mounted on 2.2°C/W Heatsink (RS components 490-7191) with silicone insulating washers
Thick lines	Represent high-current paths - up to 20A at 200W out
C3	Should be mounted very close to IC2 supply pins; IC2 should be positioned close to MOSFETs with short connections to gates and ground.
TR1	Primary connections and MOSFET source connections should be kept as short as possible

136kHz version		500kHz version	
R4, R5	39R	R4, R5	Direct connection (D4, D5 can be omitted)
R8, R9 10R,	3W metal film	R8, R9	4R7 3W metal film
C8, C9	2n2, 630V polypropylene	C8, C9	4n7, 630V polypropylene
TR1	ETD34 core and bobbin assembly, no air gap. Core material 3C85, F44, N67 or similar with mu around 2000, available from RS components, Farnell, etc. Primary 4 x 2 turns 1mm enamelled wire, each primary has 2 wires in parallel. Secondary 16 turns 1mm enamelled wire.	TR1	ETD29 core and bobbin assembly, no air gap, Core material 3C85, F44, N67 or similar with mu around 2000, available from RS components, Farnell, etc. Primary 4 x 2 turns 1mm enamelled wire, each primary has 2 wires in parallel. Secondary 16 turns 1mm enamelled wire.
C10, C12	23n; eg 22n + 1n in parallel, 1kV metallised polpropylene	C10, C12	6.2n; eg 4n7 + 1n5 in parallel, 1kV metallised polypropylene
C11	46n; eg 2x 22n + 2.2n in parallel, 1kV metallised polypropylene	C11	12.4n; eg 2x 4n7 + 2x 1n5 in parallel, 1kV metallised polypropylene
L1	Approx. 2μH - e.g. 6 turns 2 x 1mm enamelled wire on Micrometals T90-26 toroid	L1	Approx. 2μH - eg 6 turns 2 x 1mm enamelled wire on Micrometals T90-26 toroid
L2	Approx. 18μH - eg 17 turns 2 x 1mm enamelled wire on Micrometals T106-26 toroid core	L2	Approx. 7μH - eg 10 turns 2 x 1mm enamelled wire on Micrometals T90-26 toroid
L3, L4	58μH nominal - 63 turns 0.8mm enamelled wire on Micrometals T106-2 toroid core	L3, L4	15.6μH nominal - 32 turns 0.8mmm enamelled wire on Micrometals T106-2 toroid core

Table 10.3 200W class D PA component notes

improved regulation. Short, thick power cables are essential to minimise voltage drops.

The 200W PA circuit diagram is shown in Fig 10.27, with component details in Table 10.3. The push-pull current-switching type of class D output stage described in the design example above is used. It is well suited to low supply voltage, high current designs. The transverter carrier phase output signal is applied to a ‘phase splitter’ using the 74HC86 exclusive-or gates to generate antiphase square waves. These drive the gates of the output MOSFETs via the TC4427 gate driver IC. A ‘half wave’ output filter/tank circuit is used. Using the IRFP150 MOSFETs, about 85% efficiency is achieved. The MOSFETs are mounted on a 2.2°C/W heatsink attached vertically to the prototype board. For normal operation this runs quite cool, but for continuous beacon operation at full power, or if ventilation is restricted, a small fan is desirable to cool the heatsink and other output circuit components on the board. In the prototype, the air current produced by the modulator heatsink fan nearby gave plenty of cooling.

This transverter has so far been used successfully with a wide range of digital modes, including RTTY, PSK31, ‘Olivia’ and weak-signal modes such as WSPR and JT65. Also, it has been used for conventional CW and ‘visual’ modes such as QRSS and DF6NM's Chirped Hellschreiber using DL4YHF's *Spectrum Lab* software. All that is necessary to change modes is to load the appropriate software into the PC and ‘follow the instructions’. For more on these modes, see the Digital Communications chapter.

The sound card output level is set to a point just below where the HF rig ALC starts to operate. The PA modulator level can be set by setting the sound-card software to produce a CW tuning tone and adjusting the output power pot to a point just below where saturation is reached. Or better, monitor the RF output on an oscilloscope, and adjust the pot to a point just below where the modulation peaks are clipped. To operate in CW, or to generate a CW output for tuning up, etc, the HF rig is just switched to CW mode. A longer article describing this transverter system can be found at [28].

TRANSMITTER DRIVE SOURCES

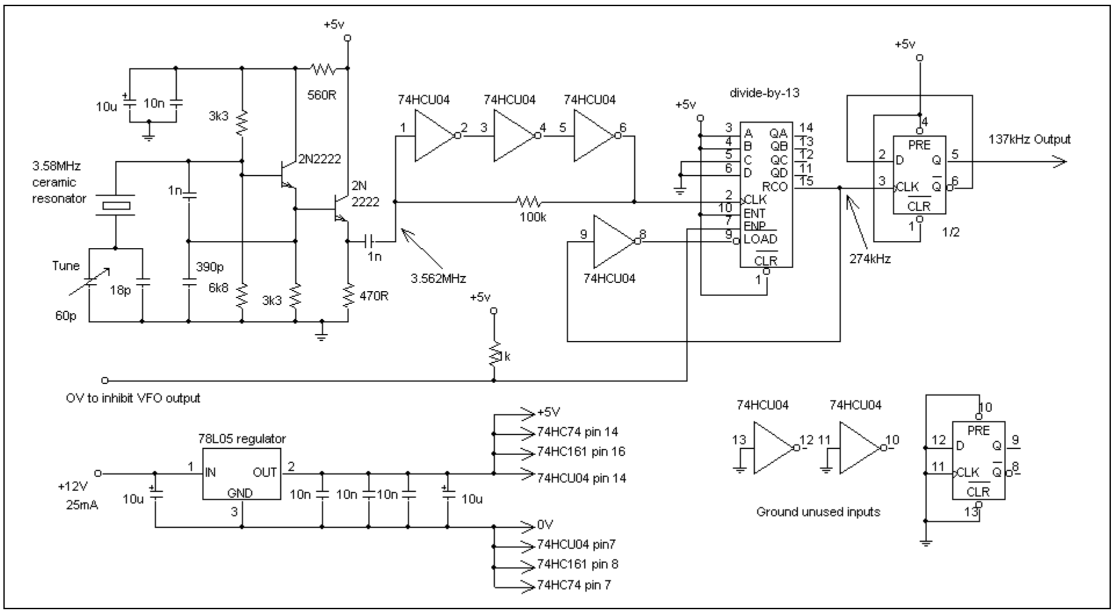
Several different types of frequency source are in use for 136kHz and 500kHz transmitters. As in the case of receivers, frequency stability requirements vary depending on operating mode. For straightforward CW operation, a simple VFO is quite adequate, the low output frequency tends to mean low drift also. Higher stability is needed for the extreme narrow band modes; frequency synthesisers of various forms are usually used. Crystal oscillators are sometimes used, usually in conjunction with a frequency divider, but the crystal frequency can usually only be pulled a few tens of hertz at the LF or MF output frequency, and being able to change frequency is very desirable even in a narrow band.

VFOs for 136kHz usually operate at higher frequencies, and are divided using digital counters to the LF range. It is easier to find suitable components for higher frequency VFOs; high stability inductors suitable for LF are difficult to produce. An example of this type of VFO is used in the G3YXM transmitter described above, where the VFO operates around 1.37MHz and is divided by 10. This is convenient for use with a frequency counter.

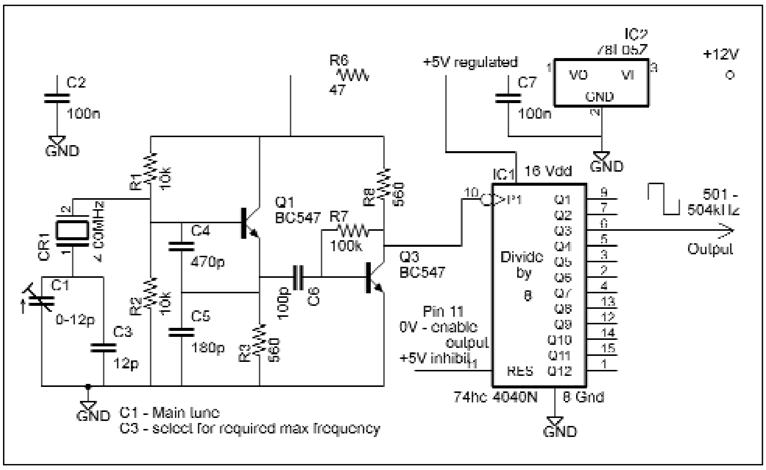
Ceramic resonators have been widely used in HF VFOs because they offer good frequency stability, whilst the frequency can be ‘pulled’ over a much wider range than crystal oscillators. Ceramic resonators are not available for 137kHz, but it is possible to use a 3.58MHz resonator in conjunction with a divide-by-26 counter to produce a VFO covering the whole 136kHz band. The complete circuit is shown in Fig 10.28.

The oscillator is pulled over the range 3.5282 - 3.5828 MHz using a 60pF series tuning capacitor and 18pF padding capacitor. The frequency tolerances of the ceramic resonators are fairly loose, so these capacitor values may need to be altered to suit individual resonators. The 1V pk-pk VFO output is buffered and then converted to a logic-level signal by an amplifier using three cascaded CMOS inverters. The signal frequency is divided by 13 by the 74HC161 programmable counter, and the resulting 274kHz signal is further divided by two by the 74HC74 flip-flop to give an

Fig 10.28: 137kHz ceramic resonator VFO



(below) Fig 10.29: 500kHz ceramic resonator VFO



accurate 50% duty cycle square wave at 137kHz. In class D transmitter designs that use a divide-by-two stage to obtain a square wave drive signal, this flip-flop can be omitted, and the 274kHz signal used directly. The resonator used has a negative temperature coefficient; the prototype circuit exhibited drift over a range of about 3Hz during a 24 hour period in an indoor environment.

Fig 10.29 shows a ceramic resonator VFO for 500kHz operation, using a 4MHz ceramic resonator with the oscillator frequency divided by eight. A similar design is used in the GW3UEP transmitter described above.

Several amateurs have used a so-called 'crystal mixer' scheme. Two crystal oscillators are operated at frequencies in the HF range, the difference in frequency being the desired output frequency. The output of the two oscillators are mixed together, and a low-pass filter at the mixer output selects the difference frequency component. Because the crystals operate at relatively high frequency, their pulling range is relatively large, and by making one or both oscillators a VXO, the whole band may be covered. A version of this system is used in the G0MRF 300W transmitter [17].

In this circuit, **Fig 10.30**, the crystal oscillators are both tuned by varicap diodes, and the oscillators and mixer are implemented using a CMOS quad NAND gate IC. This type of frequency source is less stable than a simple crystal oscillator, since drift is due to differential changes in frequency between the two oscillator frequencies, but typically the output frequency is maintained within a few hertz over a considerable period, which is adequate for many applications.

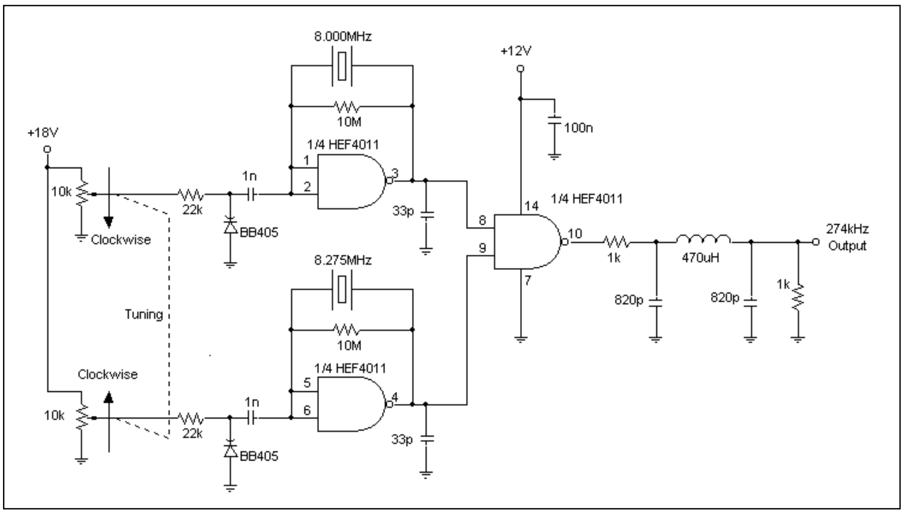
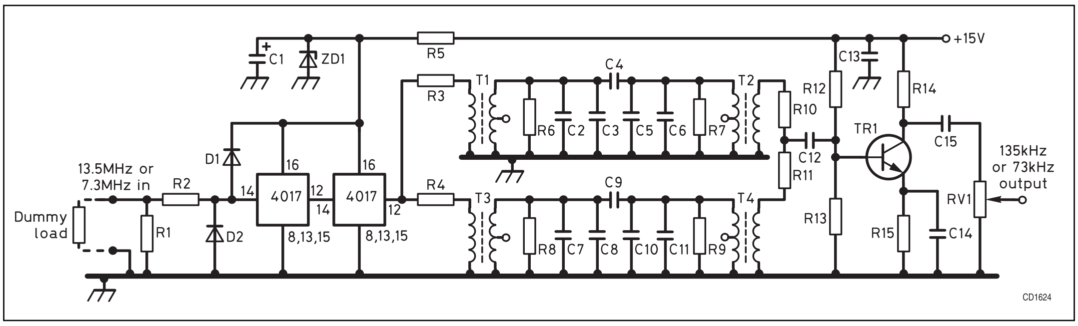


Fig 10.30: 'Crystal mixer' VFO from G0MRF transmitter

For generating modulated digital mode signals directly at LF or MF, an I/Q up-converter can be used. This reverses the signal processing steps used by receiving I/Q down-converters discussed in the Receivers section above. A modulated signal is generated at audio frequencies using PC software, with in-phase and quadrature components of the signal either being provided at the stereo sound card outputs by the software, or using hardware audio phase shift networks. The modulated audio I/Q signal is up-converted to the desired RF frequency by a fixed oscillator, with quadrature channels providing cancellation of the unwanted sideband. Tuning over a narrow range is provided by varying the audio carrier frequency in software. A

Fig 10.31: G3KAU divide-by-100 circuit



simple up-converter designed by G4JNT suitable for 136kHz or 500kHz use is described in [29] (see also the HF Transmitters chapter in this book.).

Several types of frequency synthesiser have been used as LF or MF transmitter drive sources. A number of DDS (Direct digital synthesis - see the chapter on oscillators) synthesisers have been produced suitable for 136kHz operation [30, 31, 32]. These are well suited to extreme narrow-band LF application, since they are capable of high tuning resolution, often tuning in steps of small fractions of a hertz.

Some older synthesised signal generators are available as surplus quite cheaply, especially ones that do not cover the UHF range. These often have precision reference oscillators, capable of very high frequency stability, and are obviously useful for other purposes in the shack.

Another synthesised signal source that has been widely used for 136kHz operation is an HF transceiver, whose output is digitally divided down to the LF range. A typical example of this scheme due to G3KAU is shown in **Fig 10.31**. Output from an HF rig at 13.6MHz is applied to two cascaded decade counters, giving output at 136kHz. This design also includes filtering to produce a sinusoidal output waveform for a linear transmitter. Note that most modern HF rigs inhibit transmission outside the amateur bands at frequencies such as 13.6MHz. However, in most cases only simple modification is required to enable transmission at out-of-band frequencies; contact the manufacturers for information.

TRANSMITTING ANTENNAS & MATCHING

The underlying principles of antennas are of course the same at 136kHz and 500kHz as at other frequency, but with operating wavelengths around 2200m and 600m respectively, most amateur antennas can only be a small fraction of a wavelength in length. This means they are always operated far below their self-resonant frequency, and require a large amount of inductive and/or capacitive loading in order to present a resistive load to the transmitter.

'Electrically small' antennas of this type are very inefficient, typically 0.1% efficiency would be a 'good' figure for a back-garden amateur antenna on 136kHz, rising to a few percent at 500kHz. However, in spite of this, successful amateur operation at LF and MF is done using quite ordinary wire antennas of similar dimensions to those used for HF operation.

Electrically small antennas fall into two types, vertical or loop. In the UK and Europe, the vast majority of amateurs have used vertical antennas at 136kHz and 500kHz, however LF loops have been popular with operators in North America, for reasons that will be discussed in the section on loop antennas. A well-known reference for information on amateur LF/MF antennas is ON7YD's 'Antennas for 136kHz' web page [33], which contains extensive additional information.

Vertical Antennas

The vertical or 'Marconi' antenna is the most widely used LF transmitting antenna. It consists of a vertical monopole element a small fraction of a wavelength long, driven against a ground plane, **Fig 10.32(a)**. The voltage on the element is nearly equal at all points, while the current is a maximum at the feed point, tapering to zero at the end, as the current flows to the ground plane through the distributed capacitance of the antenna element.

A figure of merit for a vertical antenna is the effective height, H_{eff}. This is the height of a notional 'ideal' vertical antenna element with a uniform current all the way along its length that would generate the same radiated field as the real antenna when fed with the same current. Because of the non-uniform current distribution, the effective height of a real antenna is always less than the physical height.

The effective height can be increased by adding top loading to the basic vertical element, as is done with the T and inverted-L shown in **Figs 10.32(b) and (c)**. These can often be existing HF dipole or long wire antennas. A large proportion of the antenna current flows to ground through the distributed capacitance of the top loading wire, increasing the current flowing in the upper part of the vertical section. In either the T or inverted-L, the distributed capacitance of the vertical section, C_V, and the horizontal section C_H is approximately:

$$C_V = \frac{24H}{\mathrm{Log}_{10}\left(\frac{1.15H}{d}\right)}, \qquad C_H = \frac{24L}{\mathrm{Log}_{10}\left(\frac{4H}{d}\right)}$$

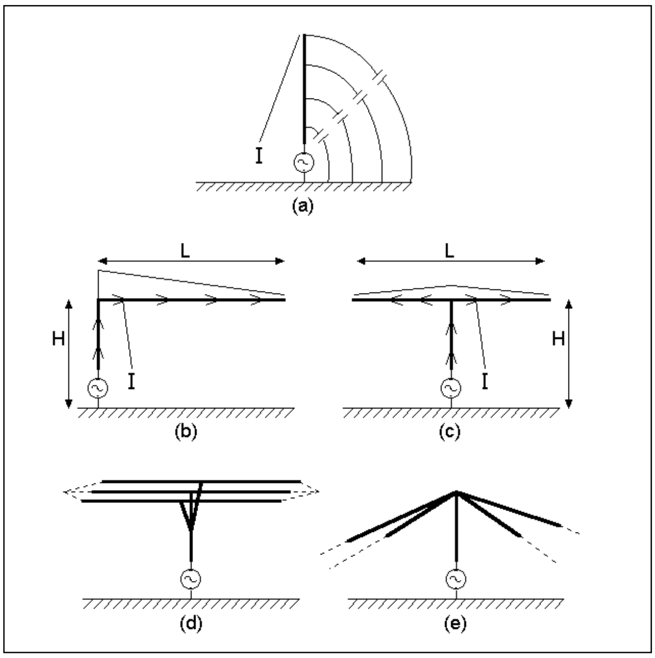
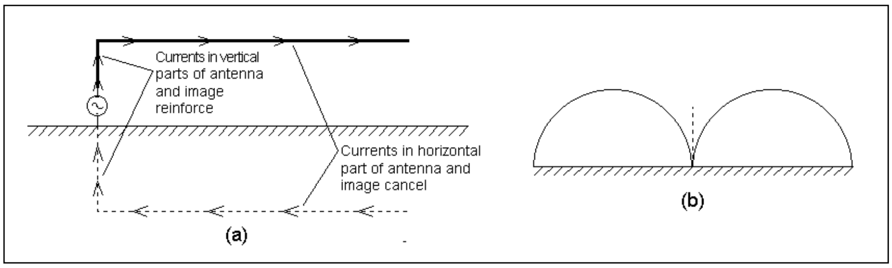


Fig 10.32: Vertical antenna configurations

Fig 10.33: (a) Cancellation of horizontally-polarised radiation from vertical. (b) Radiation pattern of short vertical antenna



Where C_V, C_H are in picofarads, H is the height of the vertical section and L the length of the horizontal section in metres, and d is the wire diameter in metres. For plastic covered wire, d can be taken as the overall diameter of the insulation. A handy approximation for C_V is 6pF per metre of height, and C_H 5pF per metre of length.

To maximise the amount of top loading in a limited amount of space, multiple top-loading wires are sometimes used, as in the 'flat-top' T antenna of **Fig 10.32(d)**. Due to proximity effects, multiple parallel wires have less capacitance than a single wire of the same total length. As a guide, two 1mm wires spaced 100mm apart will have about 39% greater capacitance than a single wire over the same span, spacing 1m apart will increase the capacitance by 68% compared to the single wire.

The effective height of the top-loaded vertical depends on the physical height H, and the relative values of C_V and C_H:

$$H_{eff} = H \frac{\frac{C_H}{C_V} + \frac{1}{2}}{\frac{C_H}{C_V} + 1}$$

With no top loading, ie with $C_H = 0$, $H_{eff} = 1/2H$. With a very large top loading capacitance, $C_H >> C_V$ and H_{eff} is nearly equal to H. Therefore, adding a large amount of top loading can nearly double the effective height of the basic vertical element. The actual shape of the top loading is not very important; the main objective is to maximise its capacitance.

Another form of vertical antenna often used at LF and MF is the umbrella, **Fig 10.32(e)**. In this case, the top loading consists of sloping wires, with the advantage that only a single tall support is required. With only two top loading wires, the umbrella becomes an inverted-V, or with one wire just a sloping long wire. The drawback is that the current flowing in the sloping wires will have a 'downwards' component, partly cancelling the 'upwards' current of the vertical section. Bringing the ends of the loading wires close to ground level is therefore likely to reduce the effective height of the antenna. Many combinations of length and angle of slope are possible, but provided the lower ends of the loading wires are at least half the height of the central vertical element, the overall effect of the umbrella top loading will be beneficial. The formulae for the T and inverted L antennas can still be used to calculate the approximate effective height of the umbrella, with the modification that H is now the average height of the sloping wires, rather than the highest vertical point of the antenna, and L is the horizontal length of the sloping wires.

Although the horizontal parts of the antenna wire are often much longer than the vertical section, little horizontally polarised radiation is generated. This is because, when the height of the horizontal section is a tiny fraction of the wavelength, the effect of the horizontally-flowing current components are almost completely cancelled out by the 'image' currents reflected in the ground plane, **Fig 10.33(a)**. Thus, these antennas are still classified as verticals, and the radiation produced is almost entirely vertically polarised. The radiation pattern of any electrically short vertical antenna is virtually the same. It is omnidirectional in the azimuth plane, and has field strength proportional to the cosine of elevation, giving rise to maximum radiation towards the horizon, and a null vertically upwards, **Fig 10.33(b)**. The directional gain of all electrically short vertical antennas is close to 2.62dB with respect to a dipole, or 1.83 as a power ratio, irrespective of their shape or size.

As an example, consider a typical antenna that might be used for 136kHz or 500kHz operation, a 40m long horizontal wire 10m above ground, made from 2mm diameter wire. Using the formulas given above, C_V = 64pF and C_H = 223pF. The antenna capacitance C_A is the sum of C_V and C_H , 287pF. H_{eff} becomes 8.9m, as expected somewhat less than the physical height.

The impedance of this and other electrically short vertical antennas can be represented by a series combination of two resistances, R_{RAD} and R_{loss}, and C_A (**Fig 10.34**). R_{RAD} is the radiation resistance, which represents the conversion of transmitter power into radiated electromagnetic waves. R_{RAD} is related to the effective height H_{eff}, and the wavelength of the radiated signal λ in metres, by the formula:

$$R_{RAD} = 160\pi^2 \frac{H_{eff}^2}{\lambda^2}$$

For 137kHz, this becomes:

$$R_{RAD} = 0.000329 \times H_{eff}^2$$

For 503kHz:

$$R_{RAD} = 0.00449 \times H_{eff}^2$$

So R_{RAD} is typically very small; for the example antenna with 8.9m effective height, R_{RAD} is only 0.026 ohms at 137kHz, and 0.36 ohms at 503kHz. The power radiated from the antenna as electromagnetic waves is I^2R_{RAD}, so quite large antenna currents are required for appreciable power to be radiated.

The other resistive component of the impedance is the loss resistance, R_{loss}, which represents the power losses in the antenna and its surroundings. These include the resistance of the antenna wires and the ground system, and power dissipated due to dielectric losses in the ground under the antenna, and in

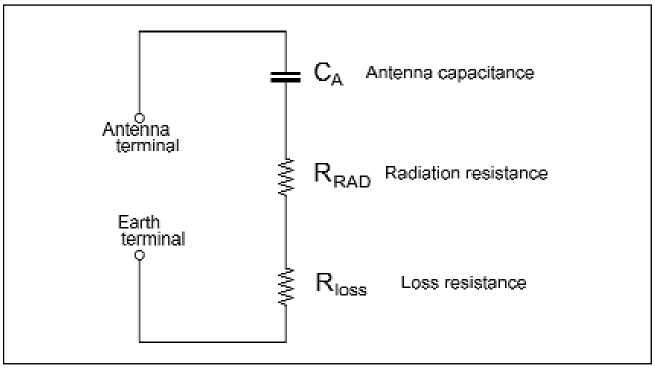


Fig 10.34: Equivalent circuit of vertical antenna

objects near the antenna, such as trees and buildings. Power dissipated in R_{loss} is converted to heat, and is therefore wasted. The same antenna current flowing in R_{rad} also has to flow through the loss resistance, resulting in a power I^2R_{loss} being wasted, so the ratio R_{rad}/R_{loss} is a measure of the antenna efficiency. R_{loss} depends greatly on the environment around the antenna.

There is no known reliable way of determining R_{loss} by theoretical calculation; measured values for typical amateur antennas in the LF and MF range vary from perhaps 10 to 200 ohms, with larger antennas having lower loss resistance, and decreasing loss resistance at higher frequency. For our example antenna an optimistic figure would be 40 ohms at 136kHz, decreasing to 20 ohms at 500kHz. The efficiency of this antenna is therefore 0.026/40 = 0.00065, or only 0.065% at 136kHz. Efficiency at 500kHz is 1.8%, nearly 30 times greater than the lower frequency, but still very low compared to antennas on the higher frequency amateur bands.

In many cases, actual efficiency is even lower, due to environmental effects; see the Practical Antenna Considerations section below.

ESTIMATING EFFECTIVE RADIATED POWER

The Schedule contained in the Amateur licence conditions specifies a maximum power level for the 136kHz band of 1W (0dBW) Effective Radiated Power (ERP), rather than the transmitter output power limit specified on most bands.

At the time of writing, at 500kHz the maximum power limit that will be considered by Ofcom is 10dBW (ie 10W) ERP. When applying to Ofcom for a Notice of Variation to operate at 500kHz, the amateur is expected to provided a justification for the power limit requested, and also to describe how the station's compliance with that limit will be determined.

It is therefore important to have an understanding of the concept of effective radiated power, and methods of determining ERP, when assembling a 136kHz or 500kHz station.

The current (2011) version of the Amateur Radio Licence Schedule defines effective radiated power as *"... the product of the power supplied to the antenna and its gain relative to a half-wave dipole in a given direction"*. The gain of the antenna can be considered to be is the product of the antenna's efficiency, and its directivity (directional gain) compared to the reference half-wave dipole. Conventionally, the dipole is considered to be in free space. The ERP can in principle therefore be determined from a knowledge of the transmitter power, antenna efficiency and directivity. Antenna efficiency can be estimated by calculating the radiation resistance R_{RAD} as described in the previous section, and measuring the loss resistance R_{loss} using an RF bridge or similar. Multiplying the transmitter output power P_{out} by the efficiency gives *the total radiated power*, P_{rad}:

$$P_{rad} = P_{out}\frac{R_{RAD}}{R_{loss}}$$

An easier way to find P_{rad} is to measure the RF current flowing at the antenna feed point, I_{ant}. The radiated power is then just:

$$P_{rad} = I_{ant}^2 R_{RAD}$$

The ERP is the radiated power multiplied by the gain with respect to a dipole, 2.62dB or a power ratio of 1.83:

$$P_{ERP} = 1.83 I_{ant}^2 R_{RAD}$$

In summary, estimating ERP involves the following steps:

- Using the formulas given previously, calculate the radiation resistance R_{RAD} using the measured dimensions of the antenna.
- Measure the antenna current (see LF/MF Measurements section for details of suitable RF ammeter designs).
- Calculate the ERP:

$$P_{ERP} = 1.83 I_{ant}^2 R_{RAD}$$

Using the previous example of a 10m high, 40m long wire antenna with radiation resistance of 0.026 ohms at 137kHz, it can be seen that an antenna current of 4.6A is required to obtain 1W ERP. The power wasted in the 40 ohm loss resistance of the antenna in producing this current is 846W, emphasising the need for large transmitter powers at LF! At 503kHz, with a radiation resistance of 0.36 ohms, a current of 3.9A is required to reach the 10W ERP limit. The transmitter power required to achieve this with 20 ohm loss resistance is only 310W.

An equivalent definition of ERP is that it is the amount of power fed to a reference antenna that would produce the same received field strength as the actual antenna does, with the conditions that both reference and actual antennas are the same distance from the receiver, and that the direction from the antennas to the receiver is the direction of maximum gain. At 136kHz and 500kHz, the reference dipole is obviously a theoretical abstraction, however it is easy to calculate what field strength E (in volts per metre) it would produce if it really existed:

$$E = 7\frac{\sqrt{P}}{d}$$

where P is the power fed to the ideal dipole and d is the distance in metres.

For example, at a distance of 10km from the dipole, with 1W feeding it, the field strength would be 700μV/m. Therefore, any transmitter and antenna combination producing a received field strength of 700μV/m at 10km distance is radiating 1W ERP. Thus, a definitive way of determining ERP is to measure the field strength at a known distance from the station. This has the advantage that it does not rely on the assumptions made regarding calculation of antenna radiation resistance and gain, but is by no means a simple measurement to make (see Field Strength Measurement section).

For antennas in ‘near-ideal locations’, ie located on open ground, free from obstructions such as buildings, trees and metal structures, it is generally found that there is close agreement between the value of ERP determined from field-strength measurements and that estimated using the antenna dimensions and antenna current. But many amateur antennas are in far from ideal locations, being in domestic gardens surrounded by many obstructions. In these conditions it is invariably found that the measured ERP is lower than the estimated value, typically by 3 - 6dB, but several decibels more for stations in urban or heavily wooded environments. This is due to reduction of the antenna's radiation resistance by the screening effects of the surroundings (see next section). It is not possible to determine this environmental loss from measurements on the antenna itself. However, since the effect of these losses is always to reduce ERP, the simple estimate above is a reliable method of setting *an upper limit* on the possible ERP of the station, and thereby ensuring compliance with licence conditions.

PRACTICAL ANTENNA CONSIDERATIONS

"As much wire as possible as high as possible" is a good guiding principle for LF and MF transmitting antennas. As seen in the previous section, the radiated power is proportional to the radiation resistance of the antenna, and the radiation resistance is

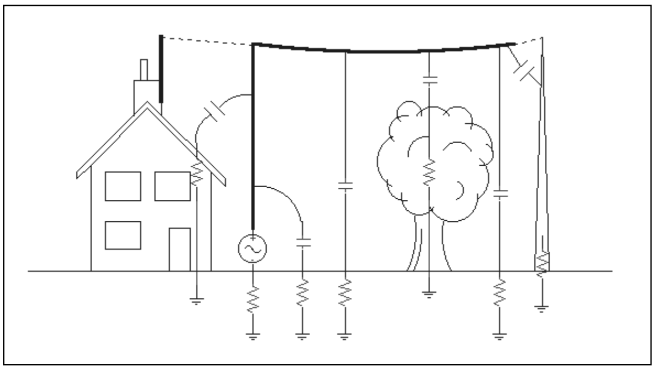


Fig 10.35: Loss in the environment around an antenna

proportional to the square of the effective height, H_{eff}. Therefore, for a given value of antenna current, and other things being equal, the amount of power radiated is proportional to the square of the height of the antenna. The most important dimension of an LF or MF antenna is therefore always its height. The effective height can also be increased by maximising the capacitance of the top loading section, so the second most important dimension is the length of wire making up the top loading section of the antenna; making this as long as possible will make the effective height of the antenna as close as possible to the physical height. However, as noted in the discussion of umbrella antennas and sloping wires, it is the average height of the top loading that is important, so having long loading wires that have a large sag, or droop close to the ground is counterproductive.

After achieving the maximum possible effective height, the next most important goal is to minimise antenna losses. The vertical antenna can be thought of as a capacitor, one plate of the capacitor being the antenna wire, and the other plate being the ground system. Losses occur due to the resistance of the 'plates', and also due to the 'dielectric', made up of the air in between, the ground underneath, and the other objects surrounding the antenna, such as buildings and trees (**Fig 10.35**). The major source of losses depends on the shape and size of the antenna. For large commercial antennas, most of the loss occurs in the resistance of the ground system, but for much smaller amateur antennas, experiments have shown that most of the loss occurs in the 'dielectric'. Part of the electric field of the antenna penetrates the ground, which is far from a perfect conductor. The electric field also induces RF currents to flow in the structure of buildings, and the wood and leaves of trees and plants, leading to further loss.

Measurements show that antenna loss resistance can be greatly increased due to these effects; the loss resistance of a particular inverted L antenna erected at M0BMU, in a domestic back garden surrounded by trees, was measured as 61 ohms at 136kHz, and 26 ohms at 502kHz, while a near-identical antenna erected in an open field for comparison had loss resistances of only 8.5 ohms at both 136kHz and 502kHz (see **Fig 10.36**).

A further environmental effect is the reduction in effective height ,H_{eff}, caused by the screening effect of the objects near the antenna. The fields close to the antenna induce RF currents flowing to ground in these surrounding objects, which therefore behave as parasitic antenna elements. Since the direction of parasitic element current flow is opposite to that in the antenna itself, and the element spacing is a small fraction of a wavelength, the radiation from the antenna is partly cancelled. Measurements show a substantial fraction of the antenna current can return to ground via these paths, significantly reducing ERP. The effective height of the antennas in Fig 10.39 calculated from the geometry of the antenna wires was 8.4m, and measurements of the radiated field strength of the open-field antenna of of **Fig 10.36(b)** yielded similar values. However, similar measurements on the home antenna of **Fig 10.36(a)** gave H_{eff} as only 5.8m at 502kHz, and 4.5m at 136kHz. Therefore the environment can be the dominant effect on overall antenna efficiency, both due to increased resistive losses and and reduced radiation resistance due to reduced effective height.

Fig 10.36: Similar antennas in different environments: (a) at M0BMU's home; (b) in open field location

Maximising the height of the antenna obviously keeps the antenna wires as far as possible from the ground and other lossy materials. Also, maximising the size of the top loading will reduce losses; a higher antenna capacitance will lead to a reduced voltage (see next section). This results in a reduced electric field intensity, reducing dielectric losses, which are proportional to the square of the field strength. The capacitance between the antenna wires and poorly-conducting objects such as trees and buildings should be kept to a minimum by keeping the antenna, including the downlead, as far from them as possible. If the antenna is supported by a metal mast, it is desirable to maximise the clearance between mast and antenna, supporting the antenna wires some metres clear using insulating halyards, or perhaps replacing the top part of the mast with fibreglass.

The effect of the mast will be reduced if it is well insulated from ground, but if it is not possible to adequately insulate the mast, ensure it has a good earth connection in order to minimise power loss due to the circulating RF current.

ANTENNA VOLTAGE AND SAFETY

An important practical consideration, particularly at 136kHz, is the antenna voltage. The RF voltage V_{ant} is almost the same at all points on the antenna, and is approximately equal to the antenna current multiplied by the reactance of the antenna capacitance:

$$V_{ant} = I_{ant} \times \frac{1}{2\pi f C_A}$$

At 137kHz, and with C in picofarads, this formula reduces to:

$$V_{ant} = 1.16 \times 10^6 \times \frac{I_{ant}}{C_A}$$

For our previous 10 metre high, 40 metre long wire example, C_{ant} = 287pF, I_{ant} = 4.6A, V_{ant} becomes 18,600V!

Very large voltages are typical, particularly when using high powers with small antennas at 136kHz. Special attention must therefore be paid to the insulation of LF antennas. At 500kHz, with reduced antenna impedance, voltages are much smaller (typically a few kV) and high voltage breakdown is much less of an issue.

A particular problem experienced by many LF operators is corona discharge. Corona occurs where the electric field around the antenna is most intense; typically at the ends of wires where they attach to insulators, at sharp bends, where loose ends of wire project, or where the antenna passes near another object that is at ground potential. Corona manifests as a continuous, diffuse electrical discharge that produces a hissing sound; it is often very hard to see, even in the dark it may appear only as a faint glow. However, it absorbs substantial RF power, and so generates a lot of heat, which can ignite plastic insulators, support ropes and antenna tuning components, leading to the collapse of the antenna. Several instances have occurred where plastic insulators, ignited by corona discharge, have dripped burning plastic on to the ground beneath.

If plastic insulators are used, ensure that they are in a position where burning debris cannot fall on a building or people. If insulators must be positioned over a building, glass or ceramic ones are safer. Even glass or ceramic insulators are eroded and cracked by corona over a period of time.

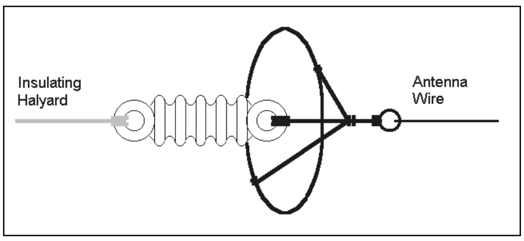


Fig 10.37: Insulator fitted with 'corona ring'

The downlead of the antenna is equally prone to corona, and this represents a particular hazard if the antenna wire is brought directly into the shack; a number of minor fires have resulted from corona discharges igniting nearby woodwork. Measures to prevent corona include the following:

- Make the capacitance of the antenna top loading as large as possible to minimise the antenna voltage.
- Use 'corona rings' (**Fig 10.37**) at ends of antenna wires, or at sharp corners, to reduce the field gradient. These can be made from loops of stiff wire, 100mm or more in diameter. Dress the ends of the wire so there are no sharp projecting points.
- Use insulating rope halyards, rather than conducting wires, to support insulators; this will reduce the voltage gradient across the insulator.
- Keep antenna wires well clear of buildings and trees, and other antennas. Locate the antenna downlead and antenna tuner away from the house or shack.

It will be seen that these are mostly the same guidelines as those given for reducing antenna losses, so taking precautions against corona also help to improve efficiency.

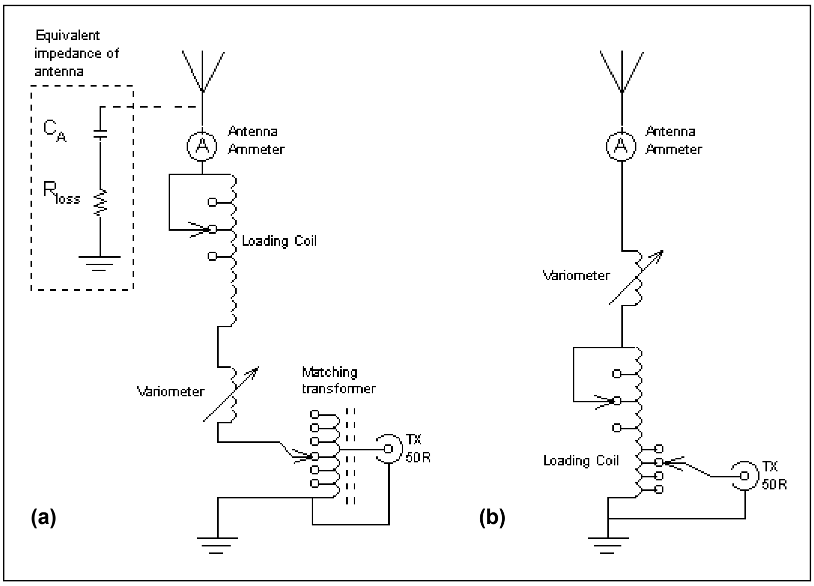
High antenna voltages are obviously also hazardous to people. Coming too close to an LF antenna wire whilst transmitting at high power can result in a severe RF burn, even without direct contact. Be sure that it is not possible for a person to come within a few metres of any part of the antenna while operating. It is also possible for large metal objects near the antenna, such as ladders, garden furniture or other antennas, to have sufficient RF voltage induced on them to cause burns to a person touching them. This can be prevented by earthing all such objects.

GROUND SYSTEMS

A ground system is required for any Marconi antenna. In professional MF, LF and VLF antenna systems, the ground system is normally the dominant source of losses, and earth mats containing many kilometres of buried wire are used. But for amateur LF and MF antennas , it is usually easy to produce a ground system that has a negligible contribution to overall antenna loss, since other losses in the antenna system are much higher.

The most widely used ground system consists of a number of ground rods connected to a common point close to the antenna tuner. This type of ground works well over much of the UK, where soil conductivity is quite high. As a very rough guide, a single 1m long ground rod will have a resistance of the order of 20 ohms; where several rods are used, spaced a few metres from each other, this figure is roughly divided by the number of rods. The losses in other parts of the antenna system are normally tens of ohms or more, so a point of diminishing returns is quickly reached where the ground system resistance is only a few ohms, and further improvements to the ground system yield little reduction in overall loss resistance. If a large number of ground rods are used, it is found that relatively little RF current flows in the rods that are further away from the feed point. This appears to be because the distributed inductance of the longer connecting wire has a large impedance compared to the rest of the ground system. This may not be true where the ground conductivity is very low. In this situation, a ground system distributed over a wider area could be expected to give a useful improvement, although little practical data is available.

Ground rods intended for domestic mains earthing are ideal; these are designed to be rigid enough to be hammered into the



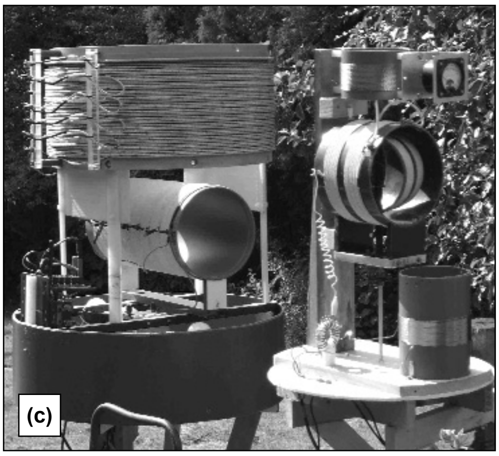


Fig 10.38: (a) LF antenna tuner; (b) Alternative antenna tuner circuit; (c) Antenna tuners at M0BMU using the circuit in (a) - 136kHz (left), 500kHz (right)

ground. It is also possible to use copper water pipe in very soft ground, or inserted into pre-made holes, but this material quickly buckles when hammered. The rods should be as long as possible, and in contact with permanently damp soil. Systems of buried radial wires as used for HF verticals have also proved effective.

MATCHING VERTICAL ANTENNAS

In principle, you could use the same matching networks used for HF antenna matching, such as the pi- or T networks, to match vertical antennas at 136kHz or 500kHz. But in practice it is found that the component values, particularly for capacitors, are impracticably large, and for 136kHz require very high ratings due to the high antenna voltage. The two most popular LF/MF antenna matching circuits are shown in Fig 10.38.

In Fig 10.38(a), a series loading coil has an inductive reactance that cancels out the capacitance, C_{ant}, of the antenna. The resistive component of the impedance (practically equal to the loss resistance) is then matched to 50 ohms, or other value of transmitter output impedance, using a ferrite-cored transformer. The capacitance of back garden amateur antennas, typically hundreds of picofarads, corresponds to a loading inductance of a few millihenries at 136kHz, and a few hundred microhenries at 500kHz.

For most antennas, R_{loss} is between perhaps 10 and 200 ohms, requiring transformer turns ratios between about 1:2 and 2:1 to match to 50 ohms. One design of 136kHz matching transformer, satisfactory at power levels up to 1kW, uses an ETD49 transformer core in 3C90 ferrite material, wound with 32 turns of 1.5mm enamelled copper wire, tapped every two turns. A much smaller transformer can be used at 500kHz, for example an ETD29 core assembly. The 50-ohm transmitter output is connected at the 16 turn tapping point, and the 'cold' end of the loading coil connected to the tap that gives optimum matching.

Because the antenna reactance is much larger than the resistance, the loading inductor must be capable of fine adjustment to obtain resonance accurately. Coarse adjustment is usually achieved by a series of taps on the loading coil. For fine tuning, the inductance is made variable over a narrow range using a variometer, described below. This matching arrangement is very straightforward to use, since the adjustment of antenna resonance and resistance loading are almost completely independent. A photo of examples of this arrangement in use at M0BMU are shown in Fig 10 38c).

Another popular matching network uses a tapped loading coil as shown in Fig 10.38(b). The low potential end of the coil is equipped with closely-spaced taps, so the loading coil also performs the function of the matching transformer. Although this is physically simpler than Fig 10.38(a), the electrical behaviour of this circuit is more complicated. The primary and secondary of the transformer are not tightly coupled, so the transformer impedance ratio will not closely correspond to the turns ratio and the adjustment of the antenna to resonance and the selection of the impedance-matching tap will be somewhat interdependent. However, it is not difficult to find a suitable tapping point by trial and error, and this will not often then need to be changed. The range of antenna loss resistance that can be matched using the tapped loading coil depends on the coil geometry. In general, if the coil has a relatively small diameter and coarse winding pitch (ie it is wound with thick wire), the maximum value of R_{loss} that can be matched is quite low. If the coil has large diameter and fine winding pitch, much higher R_{loss} can be matched. However, the coil is then less suitable for low

Fig 10.39: Antenna loss resistance matching with tapped loading coils of different diameter and winding pitch

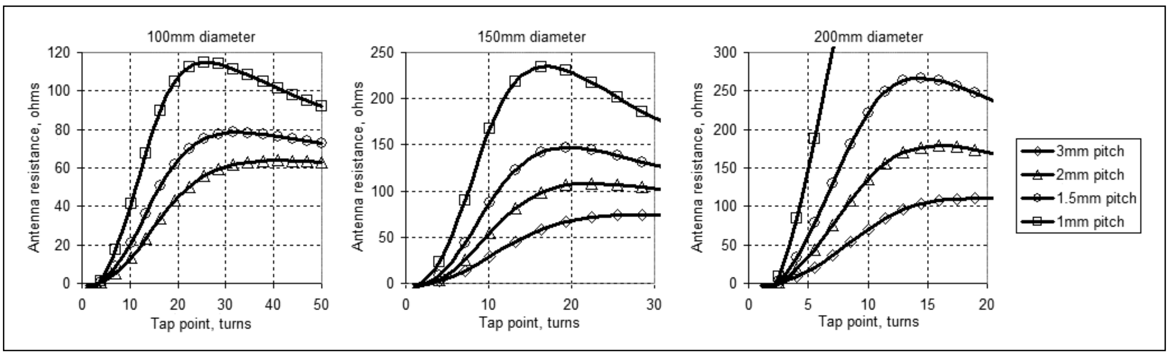


Fig 10.40: Typical loading coils

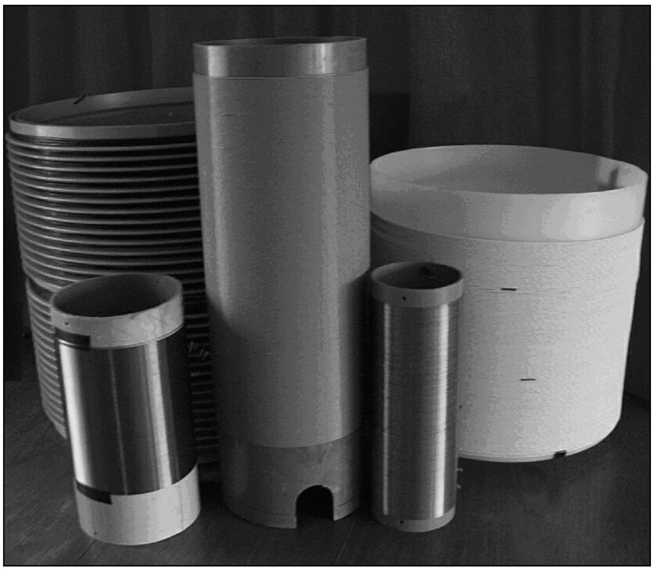


Fig 10.41: Some of the loading coils of Table 10.4

resistance antennas, because the required tap is only a few turns from the grounded end of the coil, giving very coarse steps in matching adjustment. The graphs in **Fig 10.39** show examples of the range of antenna loss resistance that can be matched with coil diameters of 100mm, 150mm and 200mm, and various winding pitches. A Microsoft *Excel* spreadsheet is available for calculating similar curves for any coil diameter and winding pitch [34].

Loading Coil Construction

The loading coil is the most critical component in the vertical antenna tuning system. It must have low losses (ie a high Q factor) if it is not to substantially reduce antenna efficiency. The coil may have to handle several amps of RF current. At 136kHz, the coil must also withstand large RF voltages. With high power 136kHz transmitters, the loading coil may need to dissipate hundreds of watts. 500kHz loading coils are less critical; smaller inductance is required, and coils consequently tend to have lower resistance. It is also necessary to make the inductance easily adjustable.

To meet these requirements, the most practical form of loading coil is a single layer solenoid of large dimensions. The former for the coil may be large diameter plastic tubing; many other roughly cylindrical plastic objects such as buckets and bins have been used to good effect. A visit to a garden or DIY centre may yield useful materials (**Fig 10.40**). Wooden coil forms have been tried, but these result in disappointingly high losses, probably due to the poor dielectric properties of wood.

Enamelled or plastic insulated wire is usually used for winding. If available, Litz wire has considerably lower loss. Litz wire is composed of many fine strands of enamelled copper wire twisted together in the form of a rope. The purpose of this construction is to reduce the 'skin effect', where at high frequencies current flow is confined to the surface of a conductor, increasing its resistance. The strands of litz wire weave in and out of the bundle, forcing current to flow throughout the thickness of the wire. This reduces the RF resistance typically by a factor of 2 or 3 at 137kHz.

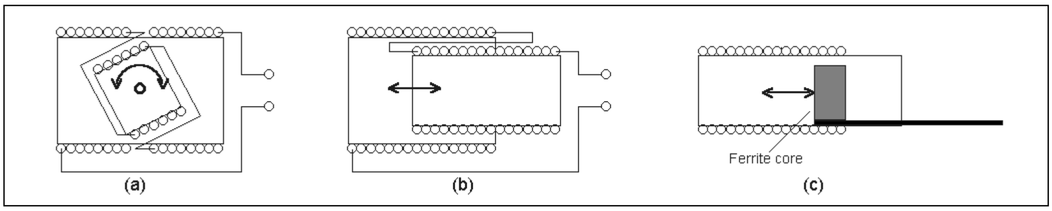
The best form of loading coil usually depends on the materials available. Data on some practical 136kHz loading coils (see **Fig 10.41**) is given in **Table 10.4**. Coils 1 and 2 are close-wound with enamelled copper wire on PVC drain pipe formers. This gives a compact coil for a given inductance, but a relatively low Q. Coil 3 is wound with teflon-insulated, stranded core equipment wire. Teflon is an excellent dielectric, and withstands high temperatures and voltages, so produces a robust, higher rated coil with lower losses. This type of wire is quite expensive, but surplus bargains sometimes appear. Coils 4 and 5 are wound with Litz wire, the 'Rolls-Royce' of loading coil materials, but hard to obtain and relatively very expensive. Coil 6 is wound with wire obtained by stripping the outer sheathing from inexpensive 2.5mm^2 'twin and earth' cable available for domestic mains wiring.

The resistance of the loading coil results in the loss of a proportion of the transmitter power in the coil. The percentage loss of power is given by:

	Former	Winding length	Wire	Turns	L, mH	Series R @137kHz	Q @ 137kHz
1	110mm dia PVC tube	280mm	0.9mm Enamelled	281	2.75	9.4	250
2	156mm dia PVC tube	190mm	0.9mm enamelled	190	3.50	15	220
3	200mm dia PVC tube.	435mm	2mm dia (1.25mm dia conductor) Teflon insulated	225	3.75	10.3	330
4	395mm dia Polythene bucket	300mm	2.7 dia 729 strand Litz	109	3.89	3.0	1100
5	485mm dia ribbed "sectional manhole", polypropylene	200mm (multi-layer)	4.1mm dia Polythene insulated 729 strand Litz	79	3.83	5.1	650
6	ditto	440mm	3.8mm dia PVC insulated	95	3.37	9.6	300

Table 10.4: Practical loading coil data

Fig 10.42: Three forms of variometer



$$\% \text{ Loss} = \frac{R_{coil}}{R_{loss} + R_{coil}} \times 100\%$$

$$\text{or, in decibels, Loss(dB)} = 10Log_{10}\left(\frac{R_{coil}}{R_{loss} + R_{coil}}\right)$$

where R_{coil} is the coil resistance, and R_{loss} is the loss resistance of the antenna.

Taking the inverted L antenna discussed above as an example, with R_{loss} of 40Ω at 136kHz, using coil 2 in the table with a Q of 220 will result in 27% of the transmitter power being dissipated in the coil, while if coil 4 with a much higher Q of 1100 is used, only 6% of the power is dissipated in the coil. The loss in radiated power in decibels is 1.4dB for coil 2 and 0.25dB for coil 4. In either case, this loss amounts to only a fraction of an S-point at the receiving station, so the effect on overall system performance is minimal.

What is more significant is the power-handling capability of the coil. Physically small coils such as 1 and 2 are suitable for transmitter power levels of up to a few hundred watts. Larger coils with higher Q dissipate a smaller proportion of the transmitter power, and also have greater surface area for cooling. Coils 3,4 and 5 are suitable for kilowatt power levels, as is coil 6, which in spite of being relatively low Q has a very large surface area.

It is wise to make the loading coil inductance somewhat higher than that calculated to resonate the antenna, and provide several taps to accommodate changes in antenna capacitance. The winding should be attached to the former in several places, otherwise there is a tendency for wire to spill off when the former expands and contacts with changes in temperature. Loading coils are often located out-of-doors, and require protection from the elements by some sort of housing.

The housing may itself cause considerable loss, especially if made of materials with high RF loss such as wood. A good protective housing is a large plastic dustbin. It is advisable to lift the housing at least a metre clear of the ground, to reduce dielectric losses in the ground underneath.

Variometers

A variometer is essentially a variable inductor, which is invaluable for fine adjustment of the loading inductance. The most common type of variometer is shown in **Fig 10.42(a)**, and consists of two coils, one rotating inside the other. When the coils are aligned so their magnetic fields add together, the mutual inductance between the coils adds to the self inductance of each coil, resulting in maximum overall inductance. Rotating the inner coil by 180 degrees results in the fields opposing, and the mutual inductance subtracting, giving minimum inductance. An adjustment range of about 2:1 in inductance is possible. Another form of variometer is shown in **Fig 10.42(b)**, in this case the two coils 'telescope' together to vary the mutual inductance.

Some examples of home-constructed variometers can be seen at [35, 36]. Reference [37] includes an on-line calculator for designing the Fig 10.42(a) form of variometer.

A very simple arrangement is shown in **Fig 10.42(c)**; this uses a ferrite core for permeability tuning of an air-cored coil. This may simply be a coil of a few hundred microhenries inductance wound on a horizontal tube former, with the ferrite core, glued to a plastic handle, free to slide inside the tube. It is important to use a short, thick piece of ferrite rather than a long rod, since in the latter case the flux density can easily become very high, leading to saturation and severe heating of the core. A large switch-mode power supply transformer core works well.

TRANSMITTING LOOP ANTENNAS

A large, single-turn loop is the 'alternative' electrically small LF transmitting antenna. While the LF vertical is a high voltage, relatively low current device, the loop features relatively low voltage, and high current. Like smaller receiving loop antennas, the transmitting loop has a figure-of-eight radiation pattern in the azimuth plane, with deep nulls at right angles to the plane of the loop. The impedance of the loop can be modelled as the loop inductance in series with a radiation resistance R_{rad} and a loss resistance R_{loss}. The value of R_{rad} depends on the area of the loop A in square metres:

$$R_{rad} = 2 \times 320\pi^4 \frac{N^2A^2}{\lambda^4}$$

where N is the number of turns, normally one. The factor of 2 is included in the formula due to the effect of the ground plane underneath the loop, which in principle doubles R_{rad} due to the 'image' antenna reflected in the ground plane.

R_{rad} for back-garden sized loop antennas, which typically have areas of hundreds of square metres, is normally below a milliohm at 136kHz. This requires loss resistance R_{loss} to be below a couple of ohms to achieve efficiency comparable with vertical antennas. The dominant source of losses for loop antennas is the AC resistance of the loop conductor. To achieve reasonable efficiency, very thick wire, or multiple parallel wires are required; several builders have resorted to using the braid of UR67 coax, or even copper water pipe!

At 500kHz, a similar loop can be expected to have a much higher radiation resistance, of the order of tens of milliohms, making efficiency comparable with vertical antennas.

Matching a loop antenna normally uses one of the circuits shown in **Fig 10.43**. **Fig 10.43(a)** uses a step-down transformer to match the loop R_{loss} to the 50-ohm transmitter output, and a series capacitance to resonate the loop inductance L_{ant}.

L_{ant} in henries is given approximately by the formula:

$$L_{ant} = 2 \times 10^{-7} P \cdot Log_e\left(\frac{3440A}{dP}\right)$$

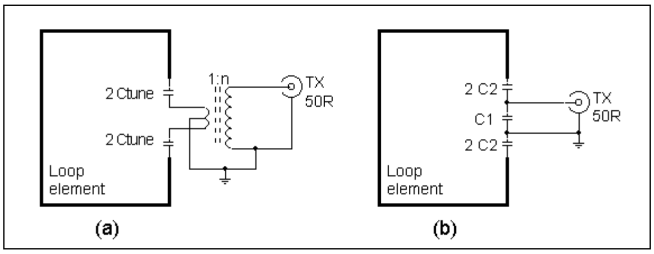


Fig 10.43: Transmitting loop matching circuits

where P is the overall length of the loop perimeter (m), A is the loop area (m^2), and d is the conductor diameter (mm). C_{tune} is therefore:

$$C_{tune} = \left(\frac{1}{2\pi f \sqrt{L_{ant}}} \right)^2$$

C_{tune} is often divided into two series capacitors as shown, to make the loop voltages approximately balanced with respect to ground. The required transformer turns ratio is $\sqrt{R_{load}/R_{loss}}$.

An alternative matching scheme uses a capacitive matching network, **Fig 10.43(b)**. The values of C1 and C2 are:

$$C_1 = \frac{\sqrt{\frac{R_{load} - R_{loss}}{R_{loss}}}}{2\pi f R_{load}}, \quad C_2 = \frac{1}{2\pi f \left(2\pi f L_{ant} - \sqrt{R_{loss}(R_{load} - R_{loss})}\right)}$$

As with vertical antennas, achieving good performance from loop antennas depends mostly on size. The radiation resistance is proportional to the square of the loop area, so every attempt should be made to make the loop as long and as high as possible. In 'open field' sites, loops are usually less efficient than similarly sized verticals for 136kHz operation due to their very low radiation resistance.

At 500kHz, a loop antenna should be competitive in efficiency with a vertical; however, at the time of writing, loop antennas do not yet appear to have been tried on this band.

The main advantage of transmitting loops is that the loop voltages are much lower than for the vertical, resulting in lower dielectric losses in objects around the antenna. This makes a loop a good choice for wooded surroundings, where many trees close to the antenna would lead to very poor efficiency with a vertical. This seems to be a common situation in North America, where several LF loop antennas have been constructed using branches of tall trees to support the loop element. Loops also do not rely on a low resistance ground connection, so may be an improvement where there is very dry or rocky soil. A disadvantage is that stronger antenna supports are required to hold the thick loop conductor. A further drawback is the directional pattern of the antenna; the radiated signal will be reduced in some directions due to the nulls in the radiation pattern and options for changing the orientation of a large loop are usually limited.

A detailed article on LF transmitting loop antennas can be found at [38]; descriptions of practical 136kHz transmitting loops are given at [39, 40].

RECEIVING ANTENNAS AND INTERFERENCE REDUCTION

Both 136kHz and 500kHz bands are subject to high levels of naturally occurring and man-made noise. The major source of naturally occurring noise is thunderstorms, which give rise to the characteristic crackling lightning static (QRN) heard at most times during the night on either band, but reduced or absent during the day.

When QRN is low, the audible 136kHz band noise floor in the UK is usually dominated by low-level sidebands from utility transmissions. These include the rhythmic chattering 'galloping horses' sound produced by the pulsed transmissions of the Loran C radio navigation system centred on 100kHz but with detectable sidebands within the 136kHz band. The sidebands of some of the high-power utility stations operating in mainland Europe and transmitting FSK data bursts in the vicinity of 136kHz are also present inside the amateur band. In contrast, the 500kHz band is relatively little affected by unwanted signals, the main one present around 504kHz possibly being the second harmonic of long wave broadcast stations on 252kHz.

A more serious problem for many amateur stations is locally generated, man-made noise. This has many sources, most associated with mains electrical wiring. Many devices use switch-mode power supplies, operating at switching frequencies in the LF/MF range. Equipment with rectifiers or triac-based phase-control circuits can generate significant levels of harmonics of the mains frequency throughout the range. Digital systems can also generate wide-band noise in the LF/MF spectrum, a particular problem if the source is a computer within the shack. Broadband internet delivered via ADSL gives rise to wide band noise that can affect the 500kHz band.

A good first stage in isolating a local noise source is to switch off all mains-operated equipment. This usually requires unplugging the equipment completely, since often the same or sometimes greater noise level can be generated when switched to 'standby'. The most certain method is to switch off the house mains supply at the main switch, whilst using a battery-operated receiver to monitor the noise level. If the offending device is on the premises, the noise can be eliminated by simply switching it off while operating. Unfortunately, mains-related noise can propagate considerable distances along the mains wiring, so often the amateur has no control over the noise source.

Many low-frequency amateur stations use their transmit antenna for reception. However, there are also many situations, especially where QRM is a problem, where it is advantageous to use a separate receiving antenna. Due to the high band noise level at low frequencies, it is possible to design very small antennas that yield perfectly satisfactory signal-to-noise ratios for reception in spite of their inevitably low efficiency.

In the case of local man-made noise, the noise level may vary greatly over short distances, and often moving an antenna only a few metres can result in substantial noise reduction. This is easily done with a compact receiving antenna where trying to move the much larger transmitting antenna to a different location is usually impractical. Many 136kHz and 500kHz operators therefore use separate transmit and receive antennas. Where the noise originates from a localised point, noise-cancelling schemes are sometimes very effective. Even if attempting to reduce the noise level is not effective, it is often found that the noise level varies a lot during the day, and is sometimes low enough for satisfactory operation. Very occasionally, the noise may abruptly disappear for good, when one noisy appliance is replaced by a quieter one, but unfortunately, the reverse is also very possible!

Receiving Loop Antennas

Much has been said about the 'noise reducing' properties of receiving loops. Electrically small loop antennas respond essentially to the magnetic field component of an electromagnetic wave, and so reject noise that exists as a local electric field. Unfortunately, most local noise sources in the LF/MF range involve common-mode noise currents flowing through mains wiring, giving rise to magnetic fields which the loop will pick up. Using a loop antenna inside or near a building therefore usually yields poor results. However, these fields rapidly decrease in strength as the antenna is moved away from the offending wiring; often moving the receive antenna by only a few metres results in substantial reduction in noise. It is therefore most important to experiment with different positions for receiving loops, often a quiet spot can be found even where high noise levels exist all around.

Loop receiving antennas have a figure-of-eight directional pattern, with nulls at right angles to the plane of the loop, and maximum sensitivity along the plane of the loop. The direction-

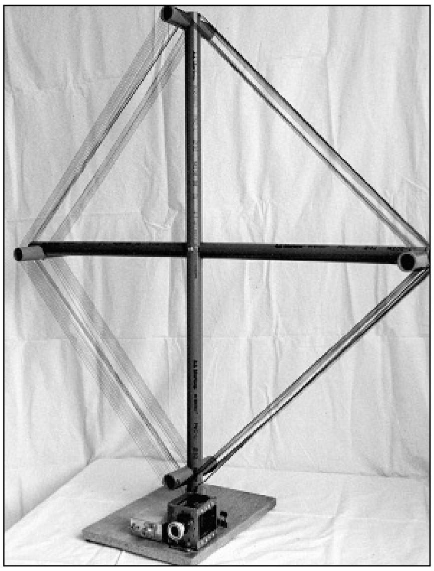
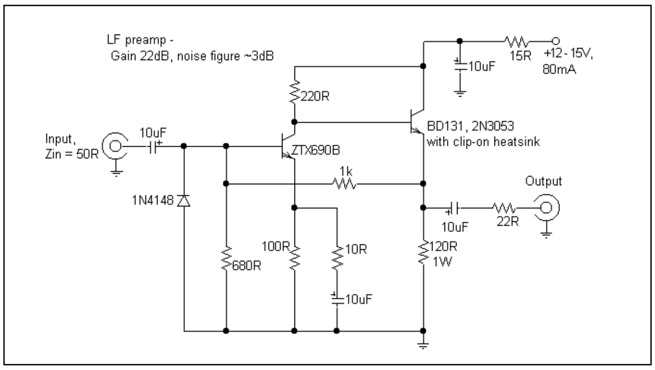


Fig 10.44: An example of a well-constructed LF loop by PA0SE



(below) Fig 10.45: 50-ohm preamp suitable for loop antennas

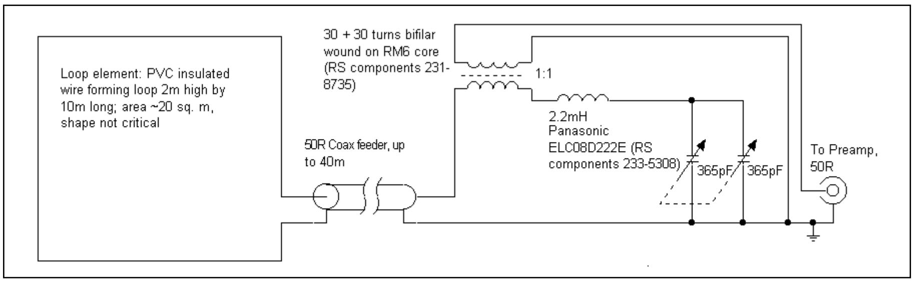
al null of a loop is often very effective in eliminating distant noise sources such as Loran. Also, the loop null can sometimes be used to suppress local noise.

A typical traditional tuned loop antenna is shown in **Fig 10.44**, and for 136kHz can consist of about 30 turns of wire wound onto a wooden cross-shaped frame, typically about 1m^2, tuned by a 1000pF variable capacitor. For 500kHz, about 10 turns is sufficient, with a 500pF tuning capacitor. The receiver input is fed via a low impedance single-turn link winding. The output of the loop is small, and a low-noise preamplifier will normally be required, such as the one shown in **Fig 10.45**. The Q of the loop is typically 100 or more, so re-tuning will be required within the narrow amateur bands. This selectivity is very useful in reducing intermodulation due to strong out-of-band signals. A number of amateurs have used much larger tuned loops for reception, which achieve higher output signal levels and so can dispense with the preamplifier, at the expense of being more bulky.

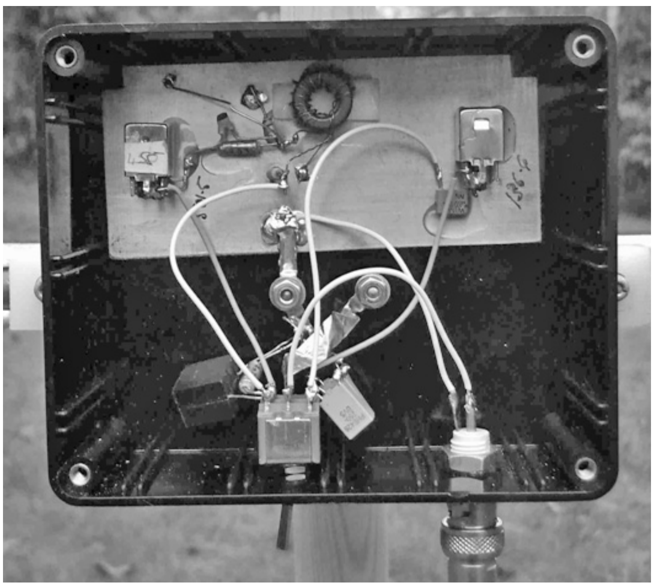
An alternative is the 'lazy loop' of **Fig 10.46**. This uses a large single-turn loop, the area of which is around 10 - 20m^2. The shape is not at all important, and it can be normal insulated wire slung from bushes or fence posts, etc., hence the name! The loop can be fed through a coax feeder, allowing it to be positioned remote from the shack to reduce noise, whilst the tuning components are easily accessible at the receiver end of the feeder. This tuning arrangement is not optimal from the point of view of minimising losses, but due to the large loop area the signal-to-noise ratio is more than adequate. It is also possible to use somewhat smaller, multi-turn loops. Again, a low-noise preamp, such as Fig 10.45 will be required. This type of loop element also gives good results with the LF Antenna tuner/preamp circuit of Fig 10.6.

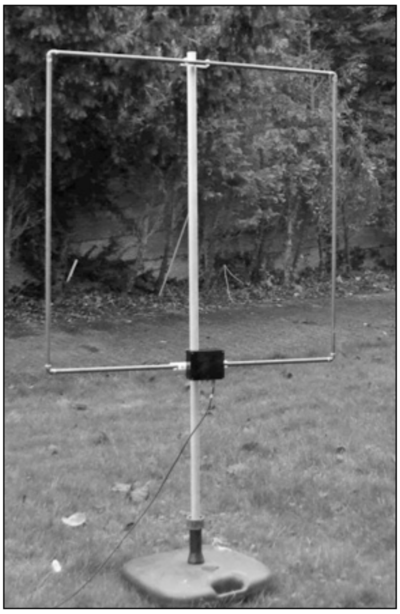
Bandpass Loops for 136kHz and 500kHz

The frequent need to re-tune a high Q loop is something of a drawback. The Q can be reduced by adding resistive loading, but unfortunately this also reduces the signal level available to the receiver. An alternative approach is to combine the tuned circuit formed by the loop with other capacitors and inductors to form a bandpass filter. This results in a wider bandwidth, while at the same time improving rejection of out-of-band signals. Another drawback of conventional



(above) Fig 10.46 'Lazy loop' and tuning arrangement





(left / right) Fig 10.47: Band-switched loop construction. The loop connections are the two bolts at the back of the case. The loop tuning capacitors C1 and C2 are mounted on the toggle band switch at the bottom

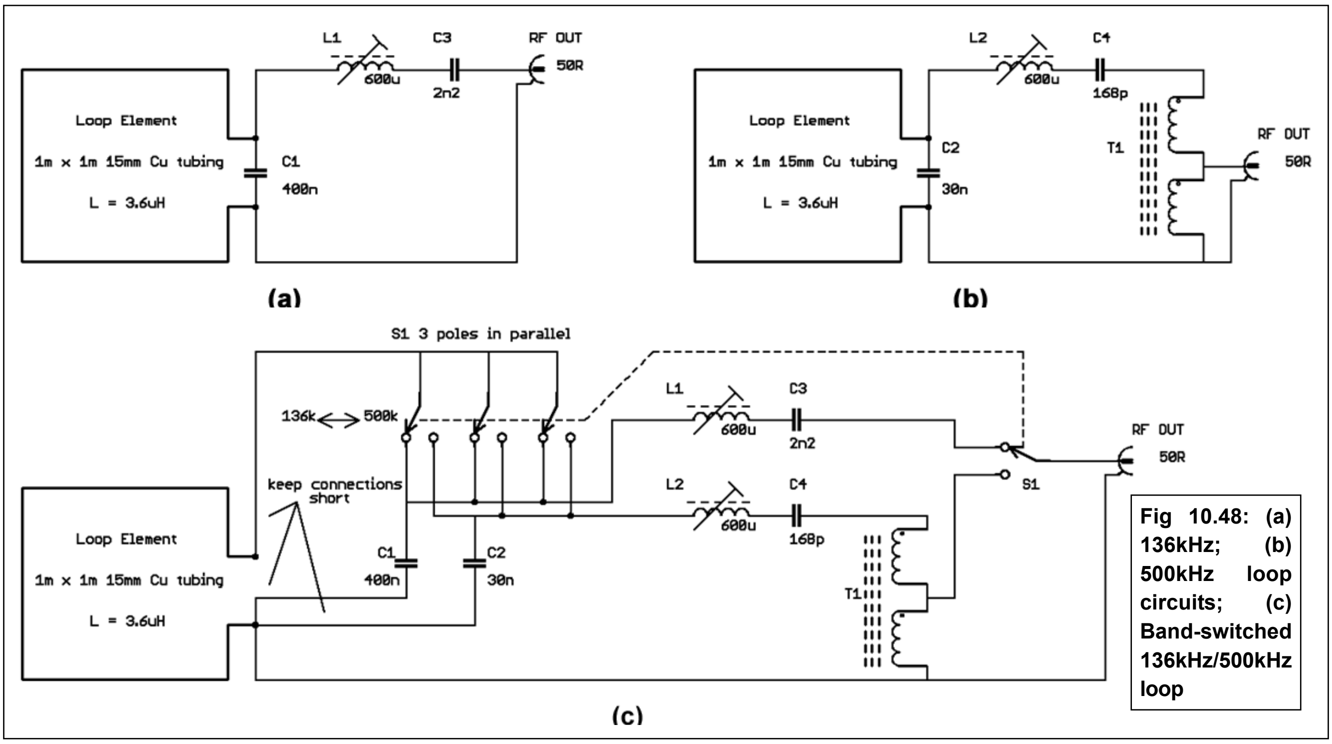


Fig 10.48: (a) 136kHz; (b) 500kHz loop circuits; (c) Band-switched 136kHz/500kHz loop

loops is the multi-turn winding, which is hard to weatherproof. A single-turn loop made of tubing is more convenient and robust, and has quite high Q, even though very large tuning capacitance is needed. The following designs are based on 1m x 1m loop elements made of 15mm copper water pipe (as shown in **Fig 10.47**) and using the bandpass filter principle to achieve coverage of the full band without re-tuning. **Figs 10.48 (a) and (b)** show similar circuits are used for both 136kHz and 500kHz loops, with the same inductance values for both. The transformer in the 500kHz design is not critical; any transformer giving a low loss and 200:50 ohm transformation at 500kHz could be used. The -3dB bandwidths of the prototypes were approximately 10kHz for the 136kHz version, and 37kHz for the 500kHz version.

A band-switched version has also been built, see **Fig 10.48(c)**. This simply switches one loop element between the two circuits. The band-change switch selecting the loop resonating capacitor must have low contact resistance in order not to increase the losses in the loop, so a 4-pole, double throw toggle switch with three poles connected in parallel was used to select the loop capacitor.

Table 10.5 shows details of the components. The 600μH adjustable inductors are the primary windings of 'Toko' or similar 455kHz IF transformers; the types that have 180pF capacitors have a suitable inductance value that is adjustable over a fairly wide range. The internal ceramic capacitor must be disconnected or removed; in the Toko types, this is most easily done by carefully breaking up the ceramic capacitor in the moulded plastic base with a pointed implement. The primary winding is usually connected to the two end pins of the row of three pins on the IFT base - check with an ohmmeter to find the largest winding resistance. Other coils with similar inductance and a Q >50 could be used.

Capacitors	Stable, low-loss types:
C1	4 x 100nF, 100V metallised polypropylene in parallel (capacitors selected during alignment)
C2	2 x 15nF polystyrene in parallel (capacitors selected during alignment)
C3	2.2nF polypropylene
C4	150pF polystyrene + 18pF in parallel
L1, L2	Primary winding of 455kHz 10mm 'Toko' IF transformer with 180pF tuning capacitor or similar; capacitor removed.
S1	4-pole double-throw miniature toggle switch
T1	18 bifilar turns on FT-50-43 (5943000301) 12.7mm dia, = 850 toroid, or similar

Table 10.5: Component notes for bandpass loops

The loop element is made from 4 x 1m lengths of 15mm copper water pipe, joined in a square with 90 degree solder elbows. One side of the loop is cut in the middle, the cut ends flattened and drilled, and brass bolts passed through and soldered into place. The bolts pass through the wall of a plastic box containing the tuning components, and connections are made using solder tags. This gives a good low-resistance connection. The connections to the loop tuning capacitors should be as short as possible and direct to the loop terminals, especially for the band-switched version - see Fig 10.47. The loop element is attached to a wooden support using plastic pipe clips.

Alignment consists of adjusting the resonant frequencies of the loop and auxiliary tuned circuits. This is best done during assembly by temporarily configuring the tuned circuits as parallel or series traps as in **Fig 10.49**; when the resonant frequency is equal to the source frequency, a sharp dip in detector level will be seen. First, the parallel resonant frequency of the loop itself is adjusted to be close to the centre of the band of interest by selecting a suitable parallel combination of capacitors (**Fig 10.49(a)**). The auxiliary tuned circuit is then adjusted to an identical series resonant frequency by itself (**Fig 10.49(b)**). Then the connections between the tuned circuits and the output are made to complete the circuit; no further adjustment should be required. Nominal resonant frequencies for the two bands are 137kHz and 504kHz, although deviations of a couple of percent from these values are not serious due to the fairly wide passband - the main thing is that both tuned circuits are set to the same frequency. Note that, when tuning the Toko inductors,

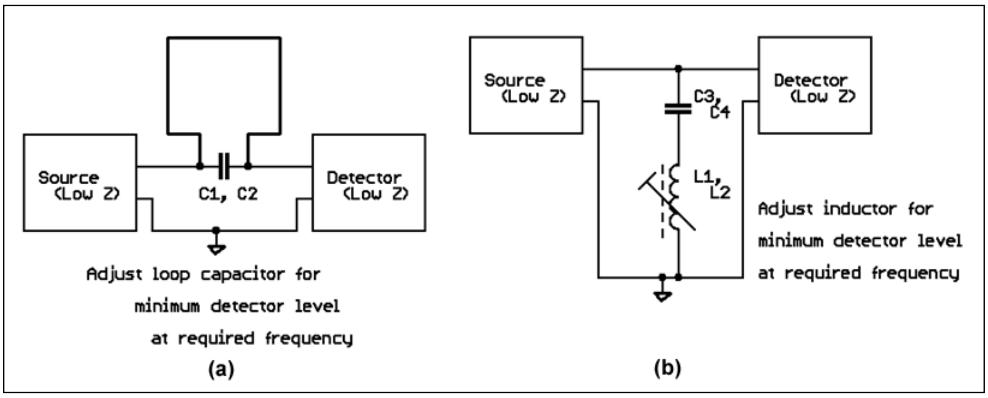


Fig 10.49: Adjusting resonant frequency of (a) loop, (b) auxiliary tuned circuit

these tiny ferrite cores can be saturated by quite low signal levels; about -20dBm from the source is safe.

The output signal level is quite small, and a low-noise pre-amplifier such as the circuit of Fig 10.45 is used. Sensitivity is then more than adequate to hear the band noise levels. The loops are not very sensitive to de-tuning; even a large metal object like a step-ladder has little effect unless it is nearly touching the loop.

These designs were originally published as a longer article which can be found at [41].

Active Whip Antennas

Quite short wire or whip antennas provide adequate signal to noise ratio at LF when matched to the receiver input by a high impedance buffer amplifier. The response of the resulting Active Whip antenna, or E-field antenna is broad-band, and can extend from the VLF range to the VHF range, depending on the amplifier used. The preamplifier is located at the base of the antenna, and its DC supply is usually fed up the coax.

The whip element is often around 1 - 2m long, and the small size makes the antenna easy to site in an electrically quiet location. As with the loop antennas, it is important to experiment with the antenna location to find a site that has a low noise level. Since all signals over a wide frequency range are presented to the buffer amplifier input, good dynamic range is important, and overload problems may occur if there are high-power broadcast stations nearby.

Unlike loop receiving antennas, the signal output level from an active whip depends strongly on its position, partly due to the screening effect of surrounding buildings and other objects, and partly on the height of the whip element above the ground plane. Greater output will be obtained if the whip is mounted on a mast, or on the roof of a building, rather than at ground level.

The PA0RDT Mini-Whip antenna

The PA0RDT-Mini-Whip© was designed by Roelof Bakker, PA0RDT, and has been built and used successfully for LF and MF reception by numerous amateurs. Good receive performance extends from 10kHz to over 20MHz. The compact size of this antenna also makes it very suitable for portable operation.

The Mini-Whip 'whip' element is in fact a small piece of copper-clad board. The shape is not important provided capacitance to ground is similar, and other whip element construction can also be used, eg a small metal box with the preamp inside. Tests were performed to optimise the size of the whip element, and the design achieves good sensitivity while maintaining maximum overall output at about -20dBm to prevent receiver overload. The buffer amplifier is optimised for good strong-signal performance. Second order output intercept has been measured by AA7U as being greater than +70dBm, and third order intercept greater than +30dBm. Power is fed from a 12 - 15V DC supply to the Mini-Whip via the power feed unit and the coaxial feed line, which can be up to 100m long. The power feed unit includes an RF isolating transformer, which reduces noise due to ground loops, but this is not always essential.

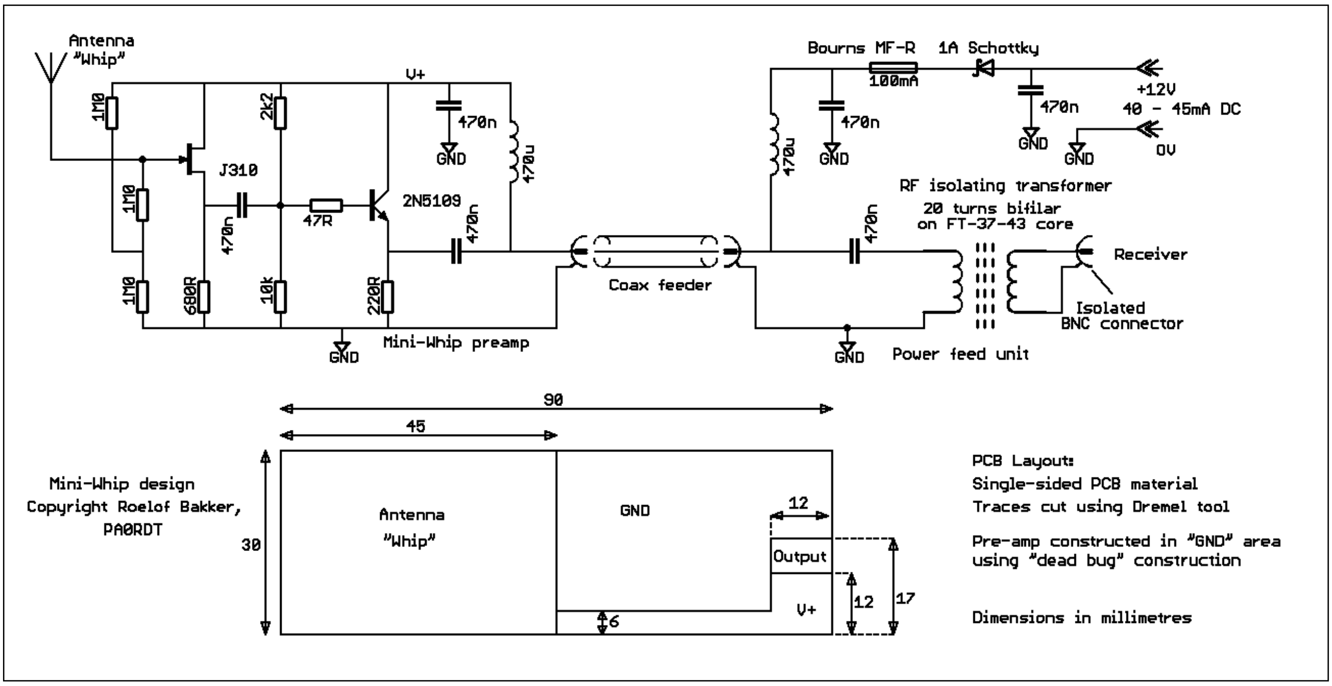


Fig 10.50: The PA0RDT-Mini-Whip©

The buffer amplifier is constructed 'dead bug' style on the ground plane half of the board next to the whip element, which is formed by the other half of the board (see Fig 10.50). The complete circuit board is mounted inside a 100mm long section of 40mm plastic drain pipe, with two end-caps. One end-cap carries an insulated BNC connector to which the circuit board is soldered.

PA0RDT notes that the electric field from most interference sources is largely confined within the building. For best results therefore, the Mini-Whip should be mounted on a non-conducting pole in a position well clear of buildings. Grounding the outer braid of the coaxial feeder to a ground rod installed close to the point where the feeder enters the shack also helps to reduce interference generated by noise currents flowing on the feeder.

A more detailed article can be found at [42]. A ready-made Mini-Whip can also be purchased from PA0RDT [43].

Terminated Loops

Terminated loop antennas such as the K9AY and EWE [44, 45] designs have been quite popular directional receiving antennas for the lower HF bands, and also are effective in the VLF, LF and MF ranges. These antennas use large wire loop elements simultaneously in a loop receiving mode and as a vertical antenna. The proper summation of loop and vertical signals at the feed point gives rise to a unidirectional pattern with a single directional null. The relative levels of loop and vertical signals can be adjusted to obtain the deepest null by varying the terminating resistor. A single loop of the K9AY type has been used successfully at M0BMU (Fig 10.51). About 20dB reduction in noise level radiated from nearby broadcast transmitters was readily obtained. It was found that a terminating resistor of around 200 ohms was required, rather lower than values quoted for the full two-loop K9AY HF array. This may be due to the use of only a single loop, or environmental factors, such as ground conductivity. Using a 1k pot as the terminating resistor will allow a wide variation to be accommodated.

The received signal level from these loops is quite low, and a preamplifier will be required with most receivers. The size and shape of the loop is not particularly critical; larger loops will generally give more output.

The Beverage antenna is another well-known form of unidirectional receiving antenna. At low frequencies, the 0.9λ length of the classical Beverage is usually impractical (about 2km at 136kHz!), but much shorter lengths have been used successfully. These probably function in a similar way to the terminated loops described above.

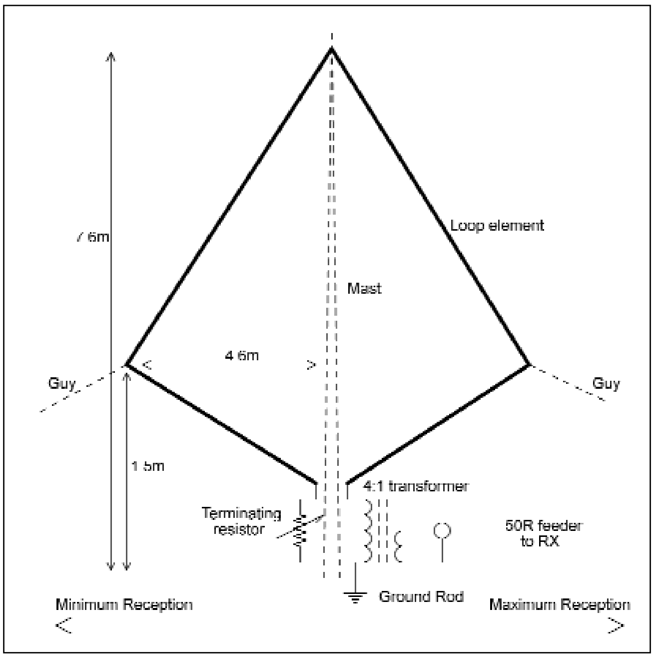


Fig 10.51: Single K9AY loop for LF/MF use

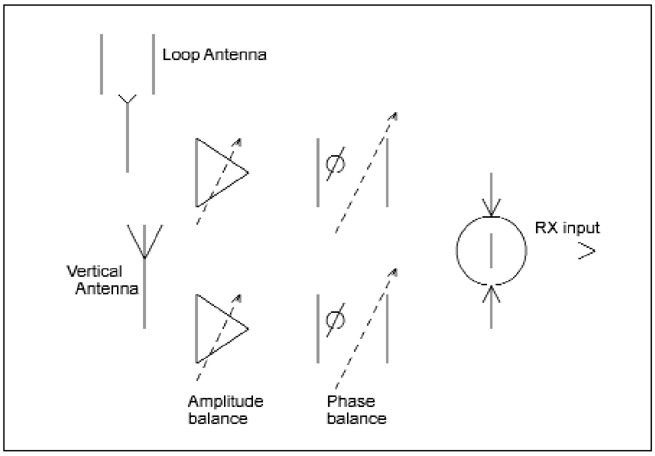


Fig 10.52: Loop/vertical directional array for low frequency reception

Noise Reduction

By combining the signals from two or more separate antennas with suitable adjustment of amplitude and phase, it is sometimes possible to cancel distant or local noise sources and so improve signal to noise ratio of wanted signals. For distant signals, this amounts to creating a directional receiving array with a null in the direction of the unwanted signal source.

Directional antenna

At low frequencies, multi-element directional arrays as used at HF and above are generally not practical due to the large spacings that would be needed between the elements. For the amateur, the most practical form of low frequency directional array consists of a rotatable loop and a vertical antenna (Fig 10.52). This type of array was widely used in the past for radio direction finding by adjusting for a null of a beacon signal, but can equally well be used to null unwanted noise.

By summing the omnidirectional vertical pattern with the figure-of-eight pattern of the loop, a skewed figure-of-eight pattern results (Fig 10.53(b)). Varying the amplitude of the vertical signal relative to the loop signal allows the skewing of the pattern to be controlled.

With zero vertical signal one obtains the basic loop pattern with two nulls at right angles to the plane of the loop (Fig 10.53(a)); as the signal from the vertical is increased, the pattern becomes asymmetrical with a smaller angle between nulls, until the limiting case of Fig 10.53(c) is reached, where the two nulls coincide to produce a cardioid pattern with a single null. Thus, the two nulls of the loop/vertical array can be 'electrically steered' to any angle between 0 and 180 degrees, whilst wanted signals in substantially different directions are received with little attenuation.

Noise cancelling

For local noise sources, noise cancelling usually relies on being able to position two antennas so that one is relatively much closer to the noise source as in Fig 10.54. Distant signals will received by both antennas at a similar level, but the 'noise' antenna will have a relatively higher level of the local noise present. Suitably attenuating and adjusting the phase of the noise

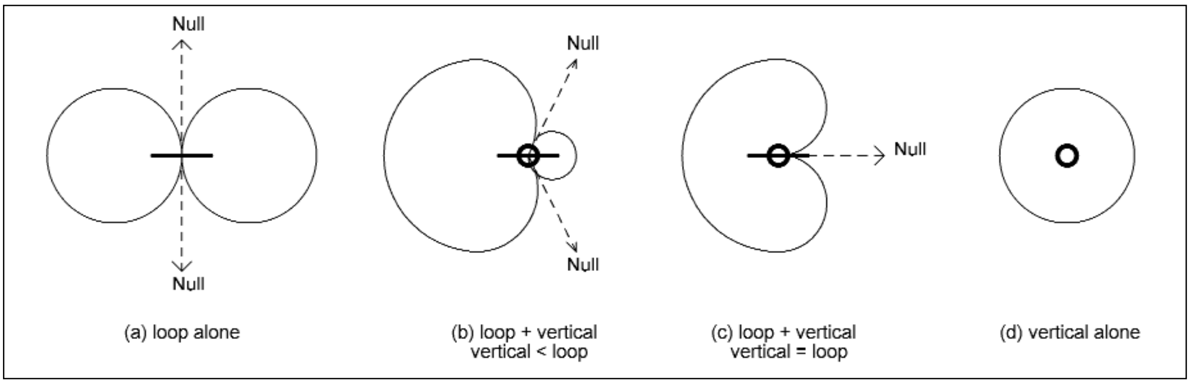


Fig 10.53: Directional patterns obtainable with loop / vertical array

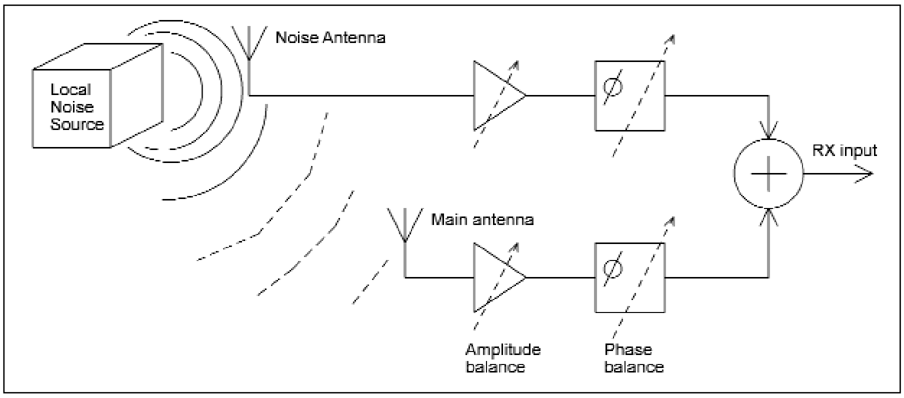


Fig 10.54: Noise-cancelling array

antenna signal before summing the outputs of both antennas results in cancellation of the local noise, with relatively little change to distant signals. To be successful, this scheme requires that the noise originates predominantly from a single source; it is unlikely to be practically possible to arrange that multiple noise sources will have the correct amplitude and phase to all be cancelled simultaneously.

It can be seen that the overall system required for obtaining directional reception or local noise cancellation is essentially the same; two receiving antennas, and a combiner that allows adjustment of relative gain and phase of the antenna signals to achieve cancellation. Very simple passive combining networks are possible, although these place restrictions on the type of antennas that can be used.

The circuit of a basic noise canceller built by G3GRO is shown in Fig 10.55. The signal from a long wire antenna feeds an adjustable phase shift network, with phase adjusted by RV1 over a range of nearly 180°. S2 is provided to allow a fixed zero degrees phase shift, since the adjustment range of RV1 does not quite reach this. Changeover switch S1 allows 180° to be added to the phase shift. The phase shifted output is simply combined in parallel with the input from a loop antenna via a variable series resistor RV2, which allows adjustment of the nulling signal amplitude. In use, the amplitude and phase controls are iteratively adjusted to achieve nulling of the noise signal. Reference [46] contains a more detailed description of this noise canceller and its use.

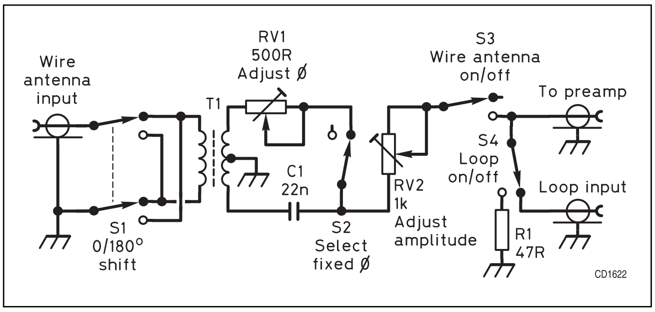


Fig 10.55: G3GRO noise canceller. RV1 and RV2 are linear cermet or carbon. T1 is 3 x 18 turns trifilar wound on a ferrite toroid (13.25mm dia. 3C85 ferrite, Amidon FT-50-43 or similar)

This simple circuit requires that the signal from the vertical antenna has a much larger amplitude than that from the loop, so that passive attenuation of the vertical signal allows the amplitudes of both signals to be adjusted to the same level and so cancel out. This could also be achieved with other types of antennas by using a preamplifier to increase the gain sufficiently.

The noise canceller can be made more versatile by providing it with adjustable gain, and phase shift that can be varied over a full 360 degree range. Buffer amplifiers can be used to isolate the gain- and phase-adjusting networks, making these adjustments independent of one another. A noise canceller of this type has been in use at M0BMU with a variety of receiving antennas at 136kHz and 500kHz. The circuit used is shown in Fig10.56.

The circuit is based on five identical high input impedance unity gain buffer circuits (Fig 10.57). Coarse adjustment of gain between -12dB and +12dB is provided at the two antenna input channels via tapping points on step-up/step down auto-transformers before being applied to the input buffers. Resistive loading of the transformers ensures a fixed input impedance close to 50 ohms independent of gain.

The input buffers each drive RC variable phase shift networks. The dual-gang 'phase balance' pot is wired differentially so that as the phase shift in one channel increases, the other channel decreases. This gives an overall phase adjustment of about +/-120 degrees. A further switched 0/180 degree phase shift is provided by inverting one channel, so that a full 360 degree range is covered with overlap.

Output from the phase shift networks is buffered and applied to a gain adjustment network. The 'amplitude balance' pot provides approximately -10dB to 0dB gain variation in each channel, and again is wired differentially so that as the pot is rotated, the signal amplitude in one channel increases while the other decreases. Together with the coarse input gain adjustment, unwanted signals differing by over 30dB at the antenna inputs can be adjusted to a null, permitting a wide range of receiving antennas to be used. Signals in the two channels are summed by connecting the gain-adjusted outputs in series, one being made floating with respect

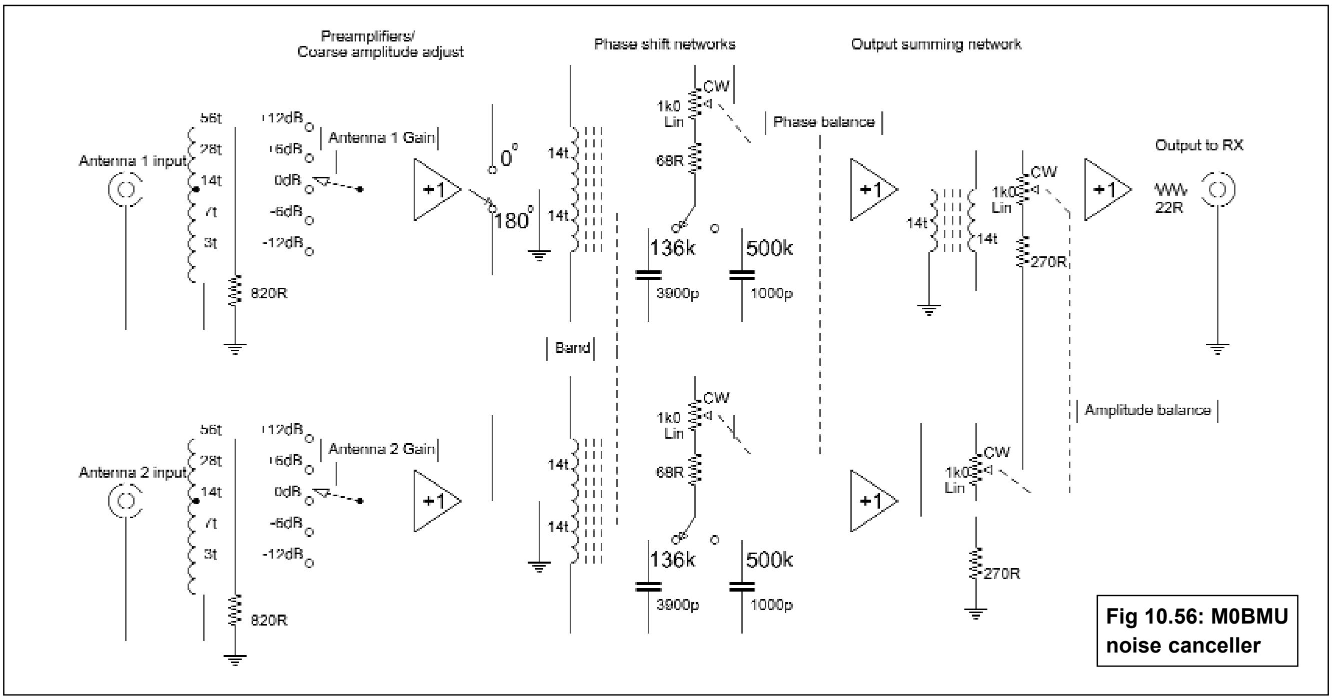


Fig 10.56: M0BMU noise canceller

to RF ground by an isolating transformer. The combined output passes through a final buffer to a low-impedance receiver input. The transformers shown in Fig 10.59 are all wound on high permeability ferrite toroids with A_L of approximately 4000nH/turn, similar to RS components part number 232-9561.

The noise canceller has been used for suppressing local noise sources, and to produce directional nulls on distant signals, mostly using the bandpass loop antennas described above, together with an un-tuned vertical antenna. Active whip and terminated loop type receiving antennas also work well. High Q resonant antennas are not well suited to noise cancelling schemes, since the signal phase and amplitude, and hence the depth of the null, is very sensitive to any alteration of the resonant frequency. Also, the bandwidth of the null is extremely small, and the controls must be re-adjusted for even a slight change in receiver frequency.

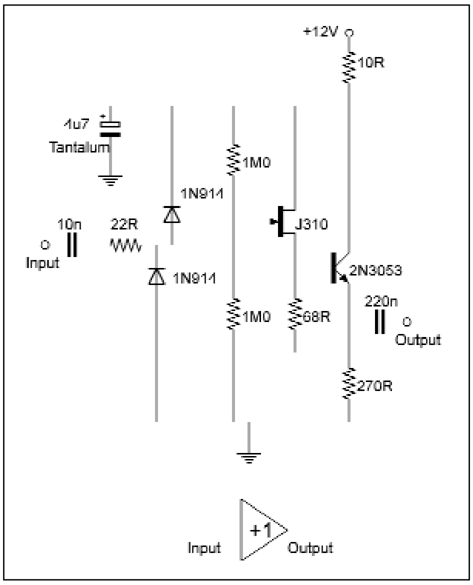
The system can be set up as follows. For nulling distant interference using the loop/vertical combination, the loop is first oriented for maximum received interference level (plane of the loop directed towards the source), or, if two sources are to be nulled, on a bearing mid-way between the interference sources. For cancelling local noise sources, one antenna is positioned to maximise noise pick-up, and the other for minimum noise. The amplitude controls of the noise canceller are then adjusted with each antenna individually connected in turn to obtain as close as possible to equal unwanted noise levels at the receiver from each antenna. Then both antennas are connected to the canceller, and the phase balance control adjusted for a null in the noise level. A few iterations of adjusting the amplitude and phase balance control should then result in a deep null.

LF MEASUREMENTS & INSTRUMENTATION

RF Ammeters

As was seen in the section on antennas, the most straightforward way of estimating the effective radiated power, or determining the efficiency of an LF station requires measurement of the antenna current. Most LF stations have some form of RF ammeter so that antenna current can be measured. The traditional RF ammeter uses a thermojunction; these are difficult to obtain, and are easily damaged by overload. Fortunately, it is easy to make a rectifier-based RF ammeter that is much more robust .

Fig 10.57: High impedance buffer used in M0BMU noise canceller



The RF ammeter of Fig 10.58(a) has a full-scale deflection of 1A, and uses a ferrite toroid as an RF current transformer. The single-turn primary is the current-carrying wire threaded through the toroid, which is wound with a 50 turn secondary. The secondary current, which is thus 1/50 times the primary current, or 20mA maximum, develops a voltage of 9.4V RMS across the 470 ohm load resistor. This voltage is measured by a simple diode voltmeter; the peak voltage across the smoothing capacitor is 1.414 x V_{RMS}, or 13.3V, less around 0.5V diode forward voltage, giving approximately 12.8V DC which the DC series resistor sets as full scale deflection of the 100µA meter. The current range can be increased by proportionally reducing the value of the load resistor; it is desirable to maintain the voltage across the load at around 10V in order to maintain the linearity of the diode voltmeter. Note that with higher maximum currents, the power dissipation in the load resistor can reach a few watts, so the resistor should be appropriately rated.

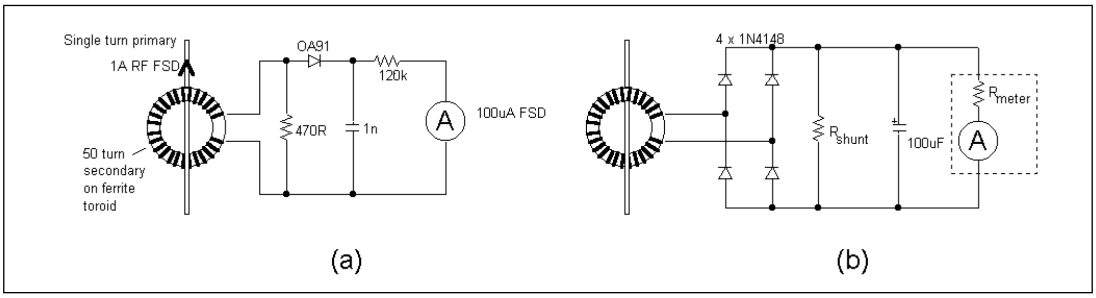


Fig 10.58: Two types of RF ammeter

A slightly different ammeter circuit is shown in Fig 10.58(b). In this circuit, the output current from the secondary of the transformer is fed directly to a bridge rectifier, whose mean output is measured by a DC milliameter. If I_{RF} is the RMS RF current, and the secondary has N turns, the mean DC current at the rectifier output is 0.90 x I_{RF}/N. The meter movement requires a shunt resistor to read the desired full-scale value. For example, if the current transformer has a 50 turn secondary, and 6A RMS full scale is required:

$$I_{mean} = 0.90 \times \frac{I_{RF}}{N} = 0.90 \times \frac{6}{50} = 0.108\text{A}$$

A 1mA meter with 75 ohm resistance was used, so the required shunt resistor R_{shunt} was:

$$R_{shunt} = R_{meter} \times \frac{I_{meter}}{(I_{mean} - I_{meter})} = 75 \times \frac{1\text{mA}}{(108\text{mA} - 1\text{mA})} = 0.70\Omega$$

The shunt resistor was made up of larger value resistors in parallel. This circuit has good linearity down to low currents because the diodes are fed from the high impedance of the transformer secondary winding.

The voltages in this circuit are essentially just the forward voltage drop of the rectifier diodes, resulting in somewhat less power dissipation than the previous circuit, and also leading to less error due to the shunting effect of the transformer inductance at lower frequencies.

A high permeability ferrite core is required for either of these circuits, so that the impedance of the secondary winding is much larger than the load impedance presented by the load/meter circuit. Toroids with a permeability of 5000 or above are ideal; with a 50 turn secondary, these typically yield inductances of several millihenries.

A split ferrite core can be used to allow the ammeter to be put into the circuit without disconnecting the wire. If the RF ammeter is to be used at a high voltage point, such as at the feed point of the antenna wire, a screening metal case enclosing the transformer and meter circuit is advisable, to prevent the possibility of stray currents flowing in the meter circuit due to capacitive coupling, and causing errors.

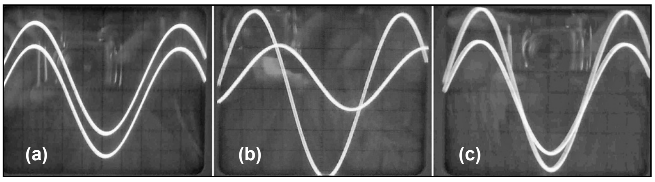


Fig 10.60: Scopematch displays - upper trace voltage, lower trace current: (a) V and I in phase and equal magnitude; load is 50Ω resistive; (b) V lags I in phase, V less than I, so capacitive load of magnitude < 50Ω (about 20Ω here); (c) V and I in phase, but V greater than I, so resistive load >50Ω (about 80Ω here)

The Scopematch Tuning Aid

136kHz or 500kHz antenna tuning is critical; the loaded Q is usually of the order of 100, giving a very sharp resonance and rapidly varying impedance as the antenna is tuned. The 'scopematch' displays the voltage and current waveforms on a dual-trace oscilloscope, which provides a convenient and intuitive visual aid to adjusting antenna tuning. If voltage and current are out of phase, the antenna impedance is partly inductive or capacitive, if in-phase the impedance is resistive. The antenna resistance can then easily be determined from the ratio of voltage to current.

The circuit diagram of the scopematch is shown in Fig 10.59. Any high permeability ferrite core of about 18mm diameter will be suitable for the current-sensing transformer, which has a single turn primary winding and 50 turn secondary. Adequate insulation is required between primary and secondary to withstand the applied RF voltage. The 1:50 transformer and 50 ohm load give a scale factor of 1V = 1A at the 'I' output, the 50:1 capacitive potential divider gives a scale factor of 1V = 50V at the 'V' output. Thus when the load resistance is 50 ohms, the amplitudes at the V and I outputs will be equal.

In use, transmitter power is applied to the antenna tuner, and the loading coil adjusted until V and I waveforms are in phase. If the amplitude of V is greater than I, loading is adjusted to reduce the resistance, or to increase the resistance if V is less than I; see Fig 10.60 for some example displays. A more detailed description of the Scopematch can be found in [47].

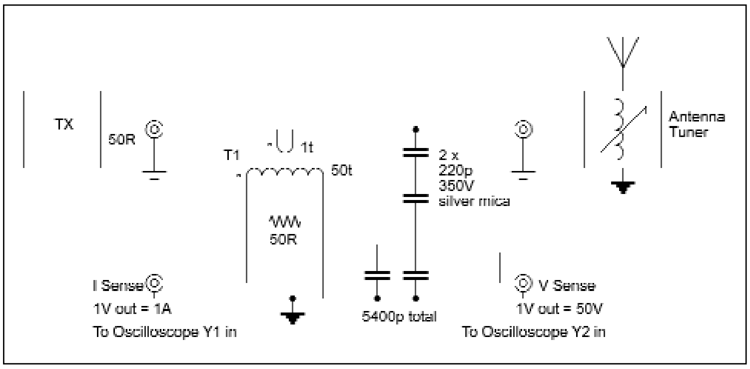


Fig 10.59: 'Scopematch' tuning aid

LF Tuning Meter

The LF tuning meter was designed as an aid to tuning transmit antennas at 136kHz and 500kHz which provides more useful information than the traditional SWR bridge. It has two meters, one indicating phase, and thus antenna resonance, while the other can be switched between measuring RF voltage and current, allowing the magnitude of the load impedance and the transmitter output power to be determined.

The circuit diagram is shown in Fig 10.61, and a picture of the prototype in Fig 10.62. The phase meter indicates the DC output of a passive double balanced diode mixer, whose inputs are the antenna current, sampled by current transformer T1, and a sample of the antenna

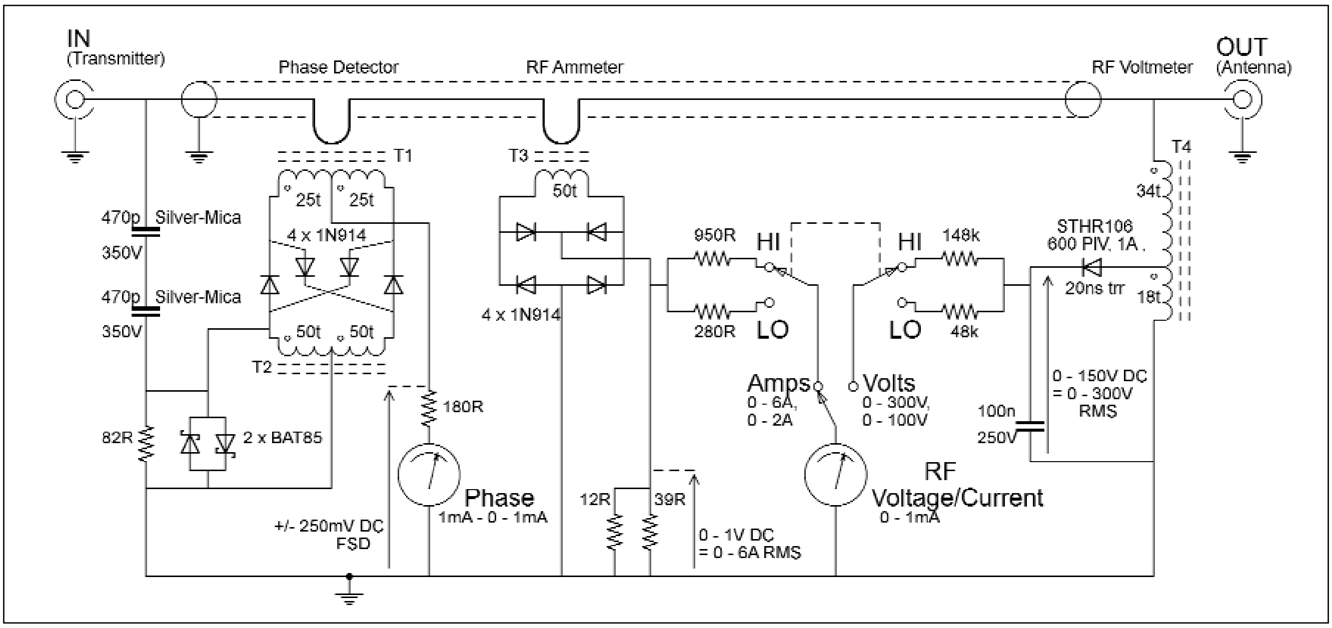


Fig 10.61: LF Tuning meter circuit. T1: Primary is RG58 coax passing through core, secondary 2 x 25t bifilar 0.3mm enamelled copper. T2: 2 x 50t bifilar 0.25mm enamelled copper. T3: Primary is RG58 coax passing through core, secondary 50t 0.5mm enamelled copper (see text for details of transformer construction)

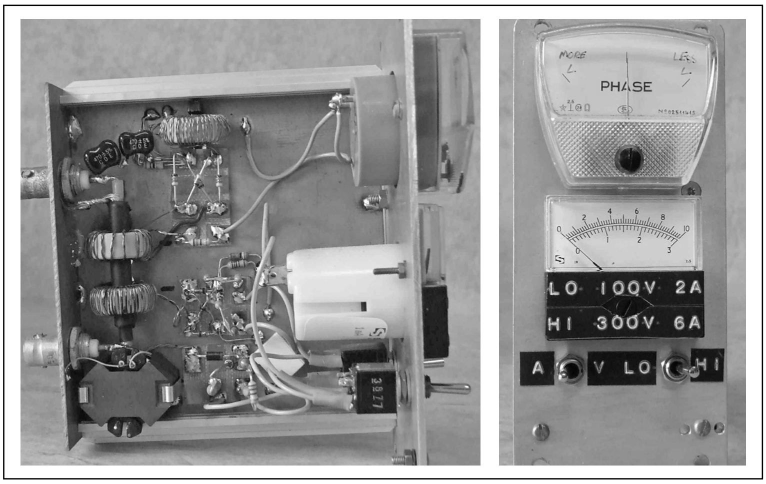


Fig 10.62: LF tuning meter prototype

voltage phase shifted by 90 degrees obtained using the potential divider formed by the series 470pF capacitors and 82 ohm resistor.

With a resistive load, antenna current and voltage are in phase, and the DC output of the mixer is zero. With an inductive or capacitive load, the phase difference between voltage and current results in a positive or negative DC output from the mixer. In this simple phase meter, output voltage depends somewhat upon power level as well as load reactance, however a resistive load gives a zero indication over a wide power range. The RF voltage is measured by a diode rectifier voltmeter driven from step-down auto-transformer T4. A toroidal current transformer T3 is used with a bridge rectifier to measure current (see RF Ammeters section above).

'Hi' (300V, 6A) and 'Lo' (100V, 2A) voltage and current ranges are provided, which makes the meter usable over a range of about 20W to 200W(Lo range) or 200W to 1.8kW(Hi range) transmitter power. The voltage and current scales are chosen so that the voltage scale is 50 times the current scale. Therefore, with a 50-ohm load connected, the same deflection is seen on the meter scale when the meter is switched between voltage and current ranges; the operator can immediately see if the load is matched to 50 ohms (V/A readings equal deflection), greater than 50 ohms (V>A), or less than 50 ohms (V<A).

The back-to-back BAT85 schottky diodes limit the phase detector input signal to a reasonably constant 500mV pk-pk with changing power levels. Other switching or small rectifier schottky diodes rated at 200mA or more will also be satisfactory. The phase detector diodes should ideally be matched to minimise the error at zero phase - this can be done using the 'diode check' range on most digital multimeters. 1N914, 1N4148 or similar diodes work well. Note the phase detector diodes are connected as a 'rat-race', while the ammeter section diodes are connected as a bridge rectifier.

The voltmeter section uses a high voltage 'ultra-fast' rectifier diode rated at 600V, 1A; similar diodes intended for SMPS applications will be suitable. The transformers T1, T2, T3 are wound on high-permeability ferrite cores about 22mm diameter with A_L of about 4000nH/t, and permeability of about 5000. The RS components 232-9561, 212-0910 and numerous others should be suitable:

The coax links the input and output sockets, and T1 and T3 are threaded onto the coax. Only one end only of the coax braid is grounded! T4 is wound on an RM10 pot core in 3C85 or similar 'power' ferrite grade. The main winding has to withstand the 300V maximum RF voltage - 'Kynar' insulated wire-wrapping wire was used instead of standard enamelled wire

The meters and range-setting resistors will depend on what you have in the junk box, and what transmitter power is to be used. Meters of different sensitivity, and different full-scale ranges can be used by changing the range-setting resistors. The values shown were assembled from series combinations during testing; presets could be used instead. Accuracy is not critical, since we are mainly just interested in the sign of the phase, and the relative level of voltage and current.

In use, select the 'Hi' or 'Lo' range to suit the power level in use. Apply transmitter power to the antenna, and note the phase meter reading. If the phase is negative (capacitive), increase the loading coil inductance, and reduce the inductance if the phase is positive (inductive). When zero phase is obtained, switch between 'V' and 'A' ranges. The ratio V/I is the resistance - if it is greater or less than 50 ohms, as indicated by the meter deflections not being equal, adjust the matching transformer, loading coil tap, etc. as appropriate. If the 'V' deflection is greater than the 'A' deflection, adjust to give a reduced load resistance; if 'V' is less than 'A', adjust for a higher load resistance. The transmitter output power when the load is matched is $V^2/50$ or 50 x I^2. A more detailed description of the tuning meter can be found in [47]

Field Strength Measurement

As pointed out in the section on antennas, the definitive way of determining the effective radiated power of a station is to measure the field strength at a known distance from the antenna. The relation between field strength E (volts/metre) at a distance d metres, and ERP is given by the formula:

$$P_{ERP} = \frac{E^2 d^2}{49}$$

For this relationship to be valid, the distance d must be in the far field region of the antenna, where the field strength falls away in inverse proportion to the distance from the antenna. The near field region is closer to the antenna, where the field strength decreases more rapidly with distance, and the formula above does not apply. For small antennas at 136kHz and 500kHz, a safe minimum distance is about 1km. At distances much greater than a few tens of kilometres, the formula also becomes invalid, due to the effects of ground loss on the propagating wave close to the ground, and ionospheric reflections. For amateur stations, the signal is also likely to be too weak to accurately measure at larger distances.

Two pieces of equipment are required to measure field strength, a calibrated receiver and a calibrated 'measuring' antenna (**Fig 10.63**). The calibrated receiver must be capable of accurately measuring signal levels down to a few microvolts, and have sufficient selectivity to reject unwanted adjacent signals; the ideal amateur equipment for this purpose is the selective level meter (see section on receivers). Calibrated antennas have a specified antenna factor (AF), which is the number of decibels which must be added to the signal voltage measured at their terminals to obtain the field strength. Quite good accuracy in the LF and MF ranges can be obtained using a simple single turn loop antenna. Such loops have a low feed point impedance, so the received signal level is little affected by the load impedance. The output voltage of an N-turn loop with area A square metres at a frequency f hertz is given by:

$$V = 2.1 \times 10^{-8} \times fNAE$$

From this, the antenna factor of a single turn loop at 137kHz is:

$$AF(dB) = 20Log_{10}\left(\frac{1}{2.1 \times 10^{-8} \times 137 \times 10^3 \times A}\right) = 20Log_{10}\left(\frac{350}{A}\right)$$

At 503kHz, AF = $20Log_{10}(94.7/A)$. A square or circular loop made of tubing is usually used, with an area between 0.5m^2 and 1m^2. As an example, suppose a signal level of 7.5dBμV (ie 7.5 decibels above 1μV, or 2.4μV; selective level meters usually give a decibel-scaled reading) is measured at a distance of 5km from the transmitting antenna, using a 1m^2 loop at 137kHz. From the formula above, AF is 51dB, so the field strength is 58.5dBμV/m, or 840μV/m. Using the ERP formula gives P_{ERP} = 350mW.

A more compact alternative to the loop is a tuned ferrite rod antenna, however this requires calibration with a known field strength to determine the antenna factor. A field strength measuring system, including ferrite rod antenna, measuring receiver, and calibration set-up has been described by PA0SE [48].

Field strength measurements are prone to errors caused by environmental factors. The measured field strength is particularly affected by conducting objects giving rise to parasitic antenna effects. Such parasitic antennas can be large steel-framed structures such as buildings and road bridges, overhead power and telephone wires, even such things as fence wires and shallow buried cables.

Such factors are difficult to avoid entirely, so several field strength measurements should be made at different locations over as wide an area as possible. Locations giving widely different values of ERP can then be rejected; it will be found that a few decibels of variation still exists between different measurement sites, so the ERP should be taken as an average of several measuring sites [49].

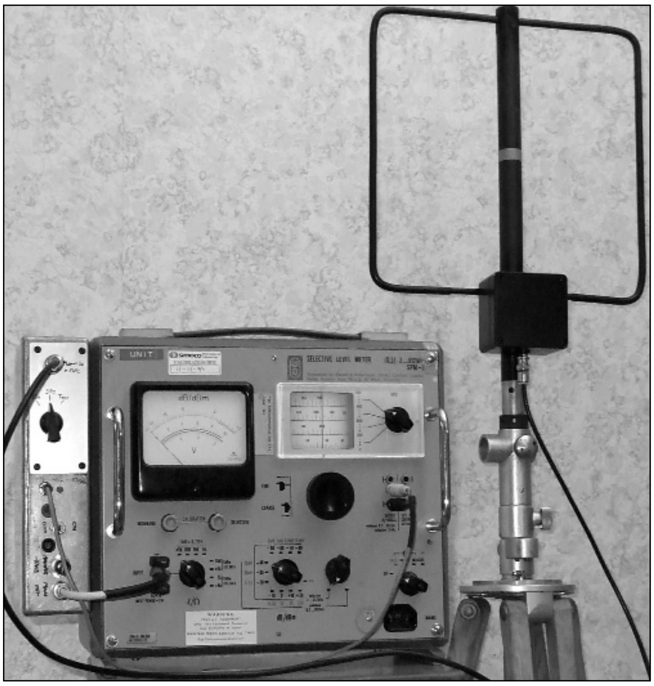


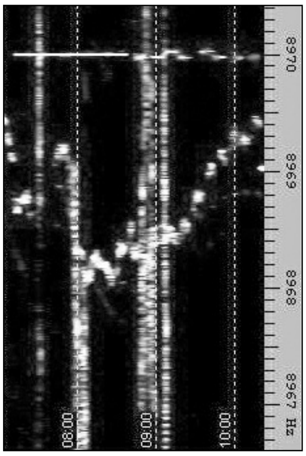
Fig 10.63: Field strength measuring system at M0BMU

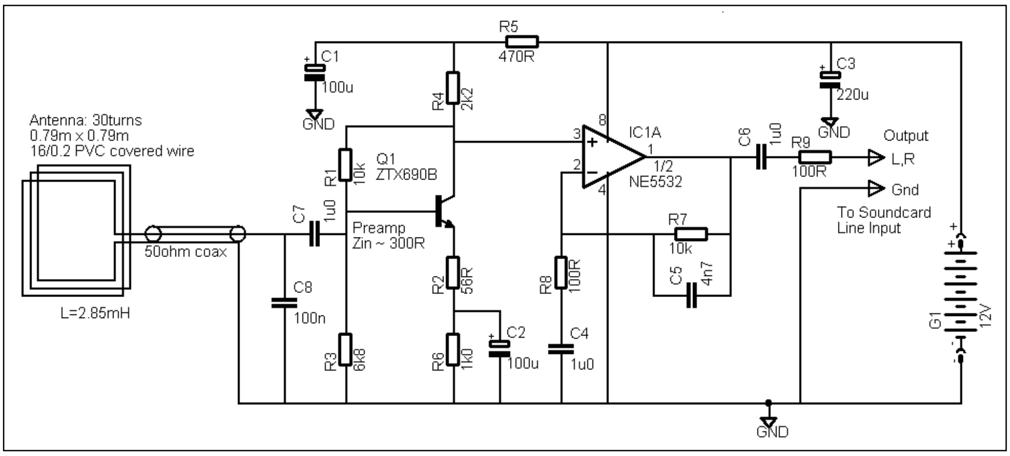
VERY LOW FREQUENCIES

The very low frequency (VLF) radio spectrum is defined as 3kHz - 30kHz. For many years it has been the nearly exclusive preserve of military communications and navigation utilities, but recently there have been several successful transmission and reception tests by amateurs in this range.

In some countries, frequencies below 9kHz are not regulated, so are freely available to amateur experimenters; in the UK, a Notice of Variation to the amateur licence must be obtained from Ofcom for transmission experiments in this frequency range, and a number of UK stations have obtained such NoVs.

After a few short-range tests, these experiments received a major boost in 2010 and 2011 thanks to the efforts of DK7FC, who carried out a highly successful series of transmissions using large kite-borne antennas and fairly high-power transmitters. Signals at 8.97kHz, 6.47kHz and 5.17kHz were widely received across Europe and as far afield as Iceland and Israel. A number of other European stations have subsequently also made successful transmissions.





(above left) Fig 10.64: 8.97kHz signal from DK7FC received over a three hour period at M0BMU, together with local noise, in February 2011. The signal consists of a fixed carrier, followed by "FB" encoded in DFCW

(above right) Fig 10.65: A VLF loop antenna together with pre-amplifier for use with an ordinary PC sound card. The input transistor, Q1 can be replaced by a BC337 with little change in performance

Transmitting and receiving techniques used at VLF are broadly similar to those used at LF, although even lower efficiencies from antennas of feasible size (the wavelength at 8.97kHz is 33.4km!) and much higher atmospheric and man-made noise levels make generating and detecting any signal very challenging.

Transmission has generally used the largest possible wire vertical antennas, with RF power generated by class D PAs, or modified audio amplifiers. Very large loading coils of the order of hundreds of millihenries are required. Efficiency is extremely low, with ERP in the micro-watt range for back-garden sized antennas, reaching milliwatts for larger kite or balloon supported antenna wires.

Signals are received using digital signal processing to generate extremely narrow-band spectrograms with milli- or even micro-hertz resolution and integration periods of minutes to hours to give a detectable signal to noise ratio. Therefore, transmissions are usually fixed-frequency carriers, or keyed with frequency shifts of a few millihertz in simple patterns to facilitate identification (Fig 10.64). Transmitting single Morse characters in this way can take hours, so communications capability is negligible, but it is possible to make technically interesting observations of propagation changes, noise, etc.

Setting up an amateur VLF transmitting station represents a major investment of time and effort, so reception reports are always very welcome. Successful reception of amateur VLF signals can be achieved with quite modest equipment, and does not require large antennas, or the obtaining of a NoV. Reception has mostly been achieved by feeding the signal from a suitable antenna / preamplifier directly into the sound card input of a PC. The standard 48kHz sound card sampling rate allows direct processing of VLF signals up to 24kHz. Most receiving stations have used DL4YHF's *Spectrum Lab* software suite [6], which performs down-conversion, filtering, noise blanking and frequency stabilisation functions, as well as generation of the spectrogram display, entirely using software. Therefore, the ordinary shack PC can be used as a remarkably sophisticated and economical software-defined radio for VLF reception.

Several types of receiving antenna have been used [50]. Large wire antennas can be effective if in a reasonably noise-free location; normally these are resonated in the VLF range (see Receive Antenna Tuning section above), and some input protection for the sound card is advisable. Small active whip and loop antennas are also successfully used, with the usual benefits of portability and the possibility of positioning to minimise local noise levels. For example, the PA0RDT active whip design described above can be modified for extended low-frequency response by increasing the values of coupling /decoupling capacitors and inductors, and increasing the input FET gate bias resistor. A simple active loop design that has been successfully used at M0BMU to receive several VLF transmissions is shown in Fig 10.65. Although response is broadly peaked at 9kHz, the bandwidth of this antenna is wide enough to receive other signals up to 24kHz, for example the occasional CW broadcasts to amateurs by Grimeton Radio SAQ on 17.2kHz.

VLF amateur radio is at an early stage in development and, although never likely to be a ‘mainstream’ activity, what has been called the ‘Dreamers' Band’ has already shown some surprising possibilities. For current up-to-date information on VLF activities, the reader is referred to the various on-line resources dedicated to this subject [51, 52].

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About the Author

__Jim Moritz__ became licensed as M0BMU in 1998, having been inspired by the introduction of the 73kHz and 136kHz bands. Most of his amateur activity has been on LF and the 500kHz experimental MF allocation. He has a background in electronics engineering. He has been actively involved in all aspects of low frequency amateur radio, including design, construction, operating and experimental investigations of antenna performance and propagation. He has also received a number of awards for his LF activities,including the Peter Bobek LF Award, the Nevada Cup and the Norman Keith Adams Prize.