HF to LF/MF Linear Transverter with 500kHz Class D

EER Power Amplifier

Jim Moritz, M0BMU, 4th November 2007

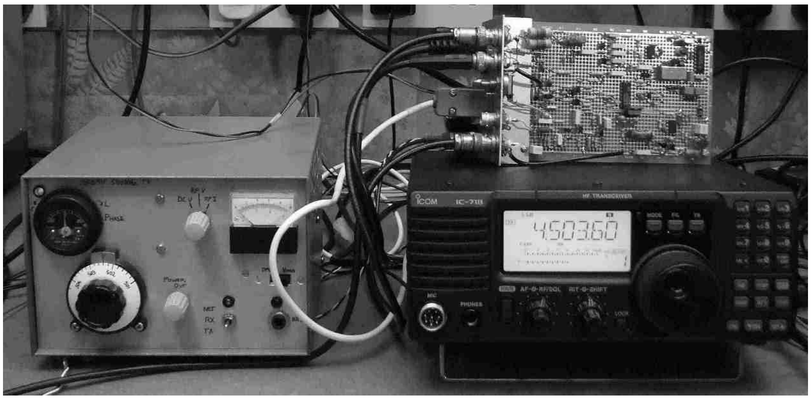


Fig 1 EER Transverter System – The transverter board is on top of the IC718 HF transceiver; the modulator and class D PA form part of the CW transmitter in the box on the left, which also contains a VFO and antenna tuning meter.

Objectives

The two main objectives of this design are:

- To provide a convenient way of transmitting and receiving digital mode signals on the 136kHz and 500kHz amateur bands.
- To act as a test bed for transmission of different modes using the EER (envelope elimination and restoration) technique for power amplification.

The signal source is an HF SSB transceiver (Icom IC718) operating at 4.0 – 4.5 MHz, 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 output in a range of about 20kHz – 550kHz. The 500kHz class D EER output stage generates about 9W PEP. For reception, the signal path is reversed to convert the LF/MF input signal up to the 4.0 – 4.5 MHz range.

EER Technique

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 modulation channel. The constant-amplitude, phase-modulated carrier frequency is applied to the PA input. 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 the amplitude 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 big, high-frequency transmitters, getting a clean signal requires complicated techniques. This is because there is often a non-linear relationship between PA supply voltage, drive and amplitude, plus unwanted phase modulation can be introduced in both modulator and PA. This can be corrected by pre-distorting the envelope and carrier phase signals using DSP techniques. But for LF/MF amateur applications at moderate power levels, class D (and presumably class E) MOSFET output stages have good supply modulation linearity, and little unwanted phase shift. Also, the modulation bandwidths are likely to be no more than a few hundred hertz, further minimising the problems with phase shifts. Requirements for spectral purity are also somewhat reduced due to low radiated powers.

A switch-mode modulator is used for EER to get the highest efficiency. But in this design, due to the low power output requirements, a simple linear modulator is used, which dissipates a significant amount of the DC input power. For types of modulation with a reasonably high crest factor (i.e. ratio between average power and PEP), which most digital modes are, the losses in the modulator are not very high. Calculating the efficiency for some common types of modulation, assuming an idealised TX system with a 100% efficient PA and a series modulator that delivers the full supply voltage to the PA at modulation peaks gives:

A1A (CW), FSK modes (no envelope modulation):	100% efficiency
BPSK modes with envelope shaping (e.g. PSK31):	89% efficiency
Two tone “IFK” modulation (e.g. “Domino”, “Throb”):	78% efficiency

The power dissipated in the linear modulator will be, as a percentage of the PEP output power:

A1A (CW), FSK modes:	Zero
BPSK modes with envelope shaping (e.g. PSK31):	7% of PEP
Two tone IFK modulation (e.g. “Domino”, “Throb”):	14% of PEP

In this theoretical system, the maximum power dissipated in the modulator is 25% of the PEP. In practice, there will obviously be some additional losses in both PA and modulator, so actual efficiencies will be perhaps 10 – 20% less than these figures for a reasonably efficient class D PA, and modulator dissipation will be somewhat higher. But this is still much better than a class AB linear PA transmitting the same modes, and for power outputs below a few hundred watts PEP, the extra complexity of a switching modulator is probably not justified in this context.

Overall System Design

A block diagram of the overall system is shown in Fig 4 (page 5). The driving HF transceiver is operated at 4.1 – 4.5MHz. 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. The transverter input circuit is wide-band, so other input frequency ranges could be used just by changing the crystal frequency. Frequencies above about 10MHz would result in poorer output frequency stability due to increased drift; frequencies below about 2MHz would require better filtering in the transverter to avoid spurious responses. 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 PA is limited to frequencies around 500kHz.

The transverter receive channel covers a similar frequency range, and achieves about 13dB SNR with a 0.1µV input signal in 250Hz bandwidth. At M0BMU, it is essential to use a separate antenna for receive, because the transmitting antenna picks up a lot of local noise in the LF/MF range. This level of sensitivity allows the receiving antenna to be quite small.

Most PC “sound card modes” software generates the modulated signal at audio frequency, which is then translated to the required carrier frequency by an SSB transmitter. To use the EER PA, the carrier phase and amplitude signals must be separated. Therefore, as well as providing the normal frequency-translating function of a transverter, this system must also be able to recover the modulation from the RF output of the transceiver. The carrier phase signal is obtained by feeding the down-converted modulated signal into a limiting amplifier. The output from the amplifier is a logic-level square wave at constant amplitude to suit the input of the MOSFET gate driver IC used in the PA. The limiting amplifier needs to provide symmetrical limiting thresholds in order not to introduce amplitude-to-phase conversion. The envelope modulation signal is obtained by rectifying the HF signal with a diode envelope detector, the output of which is 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. A fairly high level also means good linearity is obtained from the envelope detector. The detector output falls to zero before the input signal level reaches zero due to the forward voltage drop of the diodes. Although this introduces some distortion, it has the advantage that the modulator output is reduced to zero when the input to the limiting amplifier is very small, and the limiting amplifier output is noisy. This reduces the wide-band noise from the EER PA output.

The modulator/PA is based on the PA and keying/DC current limiting sections of an existing class D CW transmitter design for 500kHz. The output level of 9W PEP is sufficient to achieve the 100mW ERP “legal limit” permitted under current UK NoV regulations using the antenna at M0BMU. A buffered linear output at up to +13dBm is also available from the transverter.

Transverter Design

The schematic of the transverter is shown in Fig 5 (page 6) The prototype breadboard layout is shown in Fig 7 (page 8). The output from the driving HF transceiver is connected permanently to a 50Ω dummy load, R61, R62. On transmit, diodes D6 – D9 conduct, effectively connecting the other end of the load resistor to ground 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. The point of this circuit was to make the transverter robust against being blown up by accidentally transmitting into the receive circuits; this arrangement can tolerate a 5W level indefinitely, and 100W for short periods (until the resistors burn out!), even if the transverter DC power is not connected.

The transmit/receive diode switch uses an ICL7667 MOSFET gate driver IC (similar to TC4426) to drive about 10mA forward bias, or 2.5V reverse bias into the switching diodes D4, D5. This saves several transistor switches and bias components, and does not introduce significant noise on receive. Small signal relays could be used instead of diode switches.

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 a 74HCU04 high speed CMOS inverter. The remaining gates form a driver for the 50Ω 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/MF 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. This is necessary on transmit, where the overall gain between mixer output and PA output is high. 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 receive preamp/buffer Q1, Q9 provides a wide-band 50Ω termination at the receive input, and for the mixer and low-pass filter. The gain of this amplifier, and the overall gain of the transverter on receive, is only around 3dB. This minimises problems due to overloading by strong signals. The noise level is reasonably low to maximise sensitivity.

The transmit signal from the mixer is amplified by about 30dB by Q10, Q11. An output is taken from this point via a level-setting pot and buffer Q12, Q13 for possible use with a linear PA, or low-level test purposes. The signal is also applied to a limiting amplifier to generate a logic-level square wave output. The long-tailed pair Q3, Q4 provide a well-defined symmetrical limiting action, and Q5, Q15 bring levels up to 0V, +5V at the logic-compatible carrier phase output. This design gives a square wave output that is symmetrical within a few percent over a range of at least 30dB below the peak level of the modulation envelope.

The envelope modulation is extracted from the HF input signal by envelope detector D10, D11 and buffered by op-amp IC5A. Rectifying the HF signal results in good detector linearity due to the high signal level available. Also, 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 detector output is about +7V at the modulation peaks.

Transmit/receive switching is controlled via the PC COM port DTR handshake line for most modes. The IC718 requires its PTT line to be pulled to zero to go to transmit, which is done by Q14 (this seems to be a fairly standard rig feature … but check!). Q14 also switches MOSFET Q7 on via op-amp IC5B to produce a switched +12V TX line, which is used both within the transverter and for T/R relay switching in the PA and elsewhere if needed. This arrangement also operates the +12V TX line if the IC718 PTT line is switched to transmit internally, for instance when operating CW.

Modulator and PA design

The modulator and PA schematic is shown in Fig 6 (page 7). The prototype breadboard layout is shown in Fig 8 (page 8). The PA uses complementary N and P channel MOSFETs in a half-bridge configuration, which somewhat simplifies the gate drive arrangements. The gate driver uses a TC4427 IC driven by the carrier phase signal from the transverter (or a simple VFO for stand-alone CW operation); the R-C-Diode arrangement provides the correct DC levels, and also slows the turn-on transition of the MOSFETs to ensure both are not switched on simultaneously. The RC “snubber” components (C12, C13, C17, R14, R15, R16) help to reduce high frequency ringing on the switching transients. The PA output transformer ratio is selected to provide the required output power. The tank circuit/output filter uses two cascaded “quarter-wave” T sections, modified by the addition of C20 to produce a rejection notch at the third harmonic around 1.5MHz. The second harmonic is quite well suppressed due to the symmetrical push-pull circuit; harmonics were better than 60dB down in the prototype.

The envelope modulation signal from the transverter is applied to the reference input of the modulator, which is effectively a series regulator circuit. The feedback loop around IC3 maintains the modulator output voltage at twice the reference voltage, so the modulation input needs to peak at around 6 – 7V to get full output with a 13.8V DC supply. The pot R17 allows the modulation level to be adjusted to a level just below where clipping of the RF output signal modulation envelope occurs. Transistor Q6 provides DC current limiting in the event of a low-impedance mis-match at the PA output.

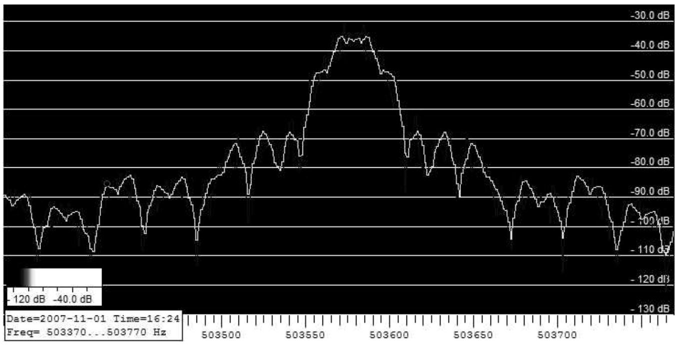
For a switching design, the combined efficiency of the PA and modulator are fairly low, at around 60 - 70% with a 9W CW output. Most of the losses are due to the on-resistance of the IRF520 and especially IRF9520 devices. More modern MOSFETs are available with much lower $R_{DS(on)}$, but these devices were used because they were to hand, and the output power required was low.

The schematic in Fig 5 also includes the receiver input bandpass filter, which provides useful rejection of high-power MF and LF broadcast signals for the broadband transverter RF input. The filter bandwidth is about 150kHz, and does not require tuning. The antenna relay is operated by the +12V TX line from the transverter. The receive filter input can be connected to the transmit antenna via the relay, or to a separate receive antenna. If used, the receive antenna should have its own T/R switching arrangements to restrict the amount of RF power reaching the receiver input during transmit, and to reduce possible feedback to the input of the transmit amplifier chain.

Results and Conclusions

This transverter has so far been tried successfully with RTTY, PSK31 and DF6NM's Chirped Hellschreiber modes. All that is necessary to change modes is to load the appropriate software into the PC and "follow the instructions". 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 generate a CW output for tuning up, etc, the HF rig is just switched to CW mode. The 500kHz signal output seems to be quite clean; Fig 2 shows the close-in spectrum of a PSK31 signal, showing all unwanted sidebands suppressed by >30dB. Fig 3 is a wider bandwidth spectrum plot, showing that most of the noise in the output occurs within the HF rig SSB filter bandwidth; at least some of this is noise originating from the PC sound card output.

A previous EER system at M0BMU used a similar Class D PA / linear modulator scheme to transmit 1.2kW PEP BPSK signals on the 136kHz band. This worked well, but was limited in its application because it used dedicated waveform generating hardware to produce amplitude modulation envelope and phase keying signals, which was only designed for BPSK modes. It also required the raw binary data to do this, which is not available from most "sound card modes" software. The system described above is much more flexible and produces an adequate quality of signal. Technically, it is not the ultimate – the multiple frequency conversions involved, and the many points at which noise and distortion can be introduced are bound to degrade the signal to some extent on both transmit and receive. However, it does allow interesting modes to be tested using available software, and demonstrates the practicality of the EER Technique.



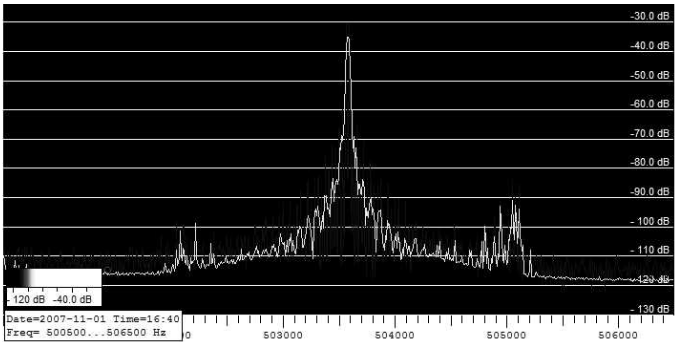


Fig 2 Close-in spectrum (+/- 250Hz) of PSK31 Signal

Fig 3 Wider bandwidth spectrum (+/- 3kHz) of Chirped Hell signal

Future additions and modifications under development include:

- Addition of a high power EER PA for 136kHz operation, probably with a switching modulator.
- Deriving the 4MHz conversion oscillator signal from the frequency standard inside the HF rig – this will improve frequency stability to better than 1ppm for use with specialised narrow-band modes such as "Wolf" PSK.

Version 1

Version 2 – Added photographs and some minor corrections 11/11/2007

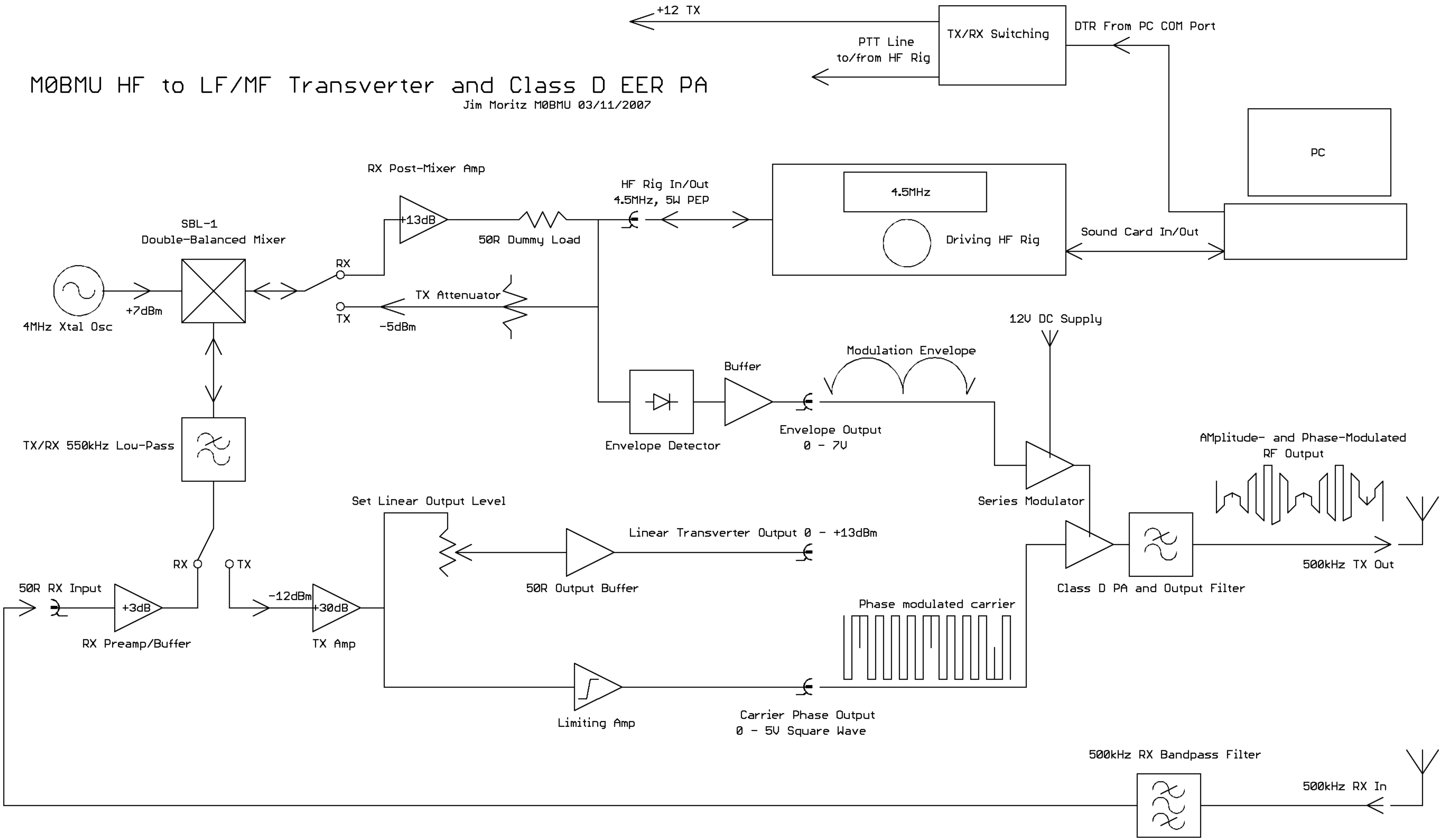


Fig 4 – Block Diagram of LF/MF Transverter and EER PA

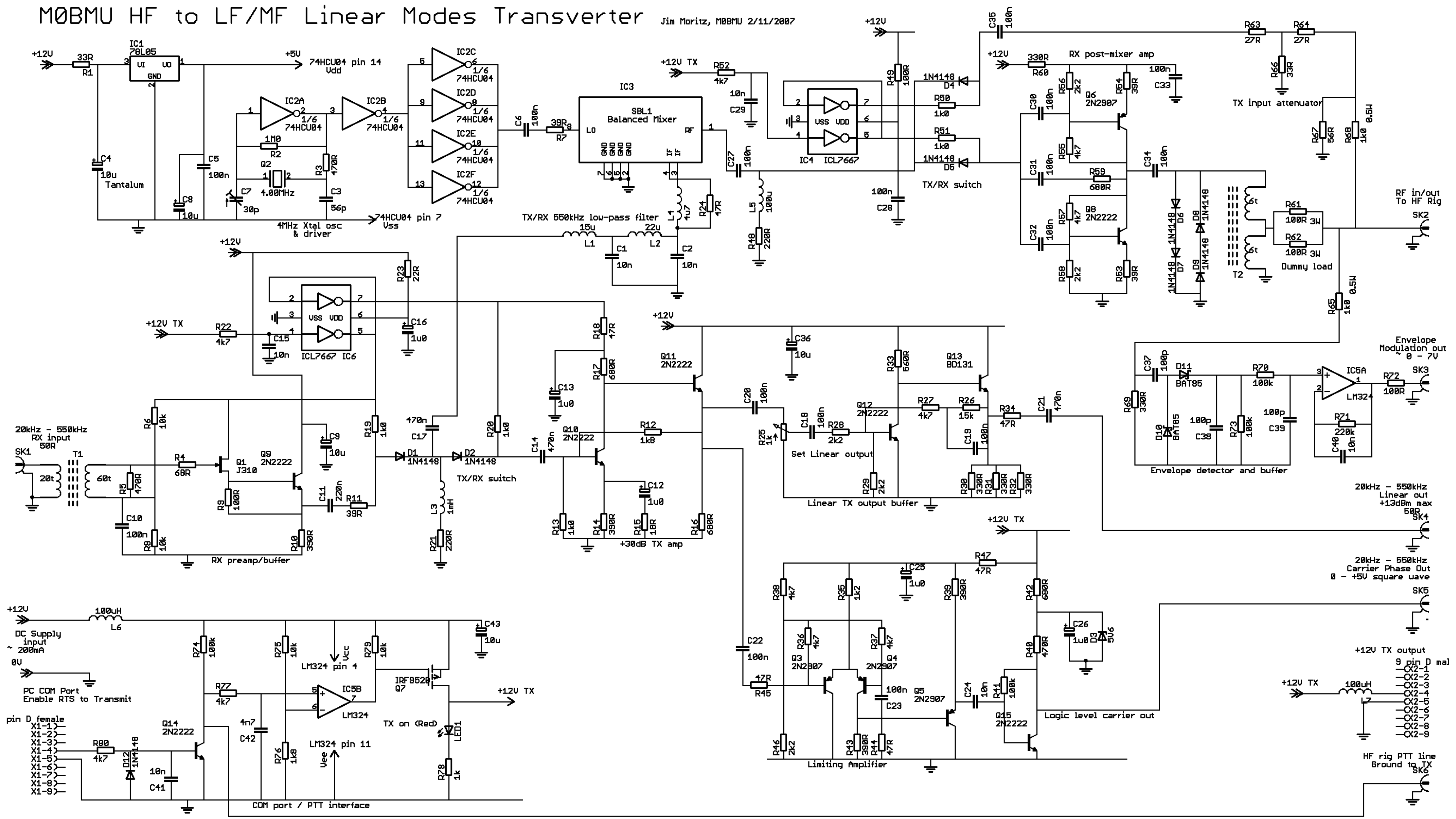


Fig 5 Transverter Schematic

T1 - 20 turns primary, 60 turns secondary on 14mm dia 3C85 toroid core
T2 – 2 x 6 turns 0.4mm wire bifilar wound on 14mm dia 3C85 toroid core
Electrolytic capacitors are tantalum types.

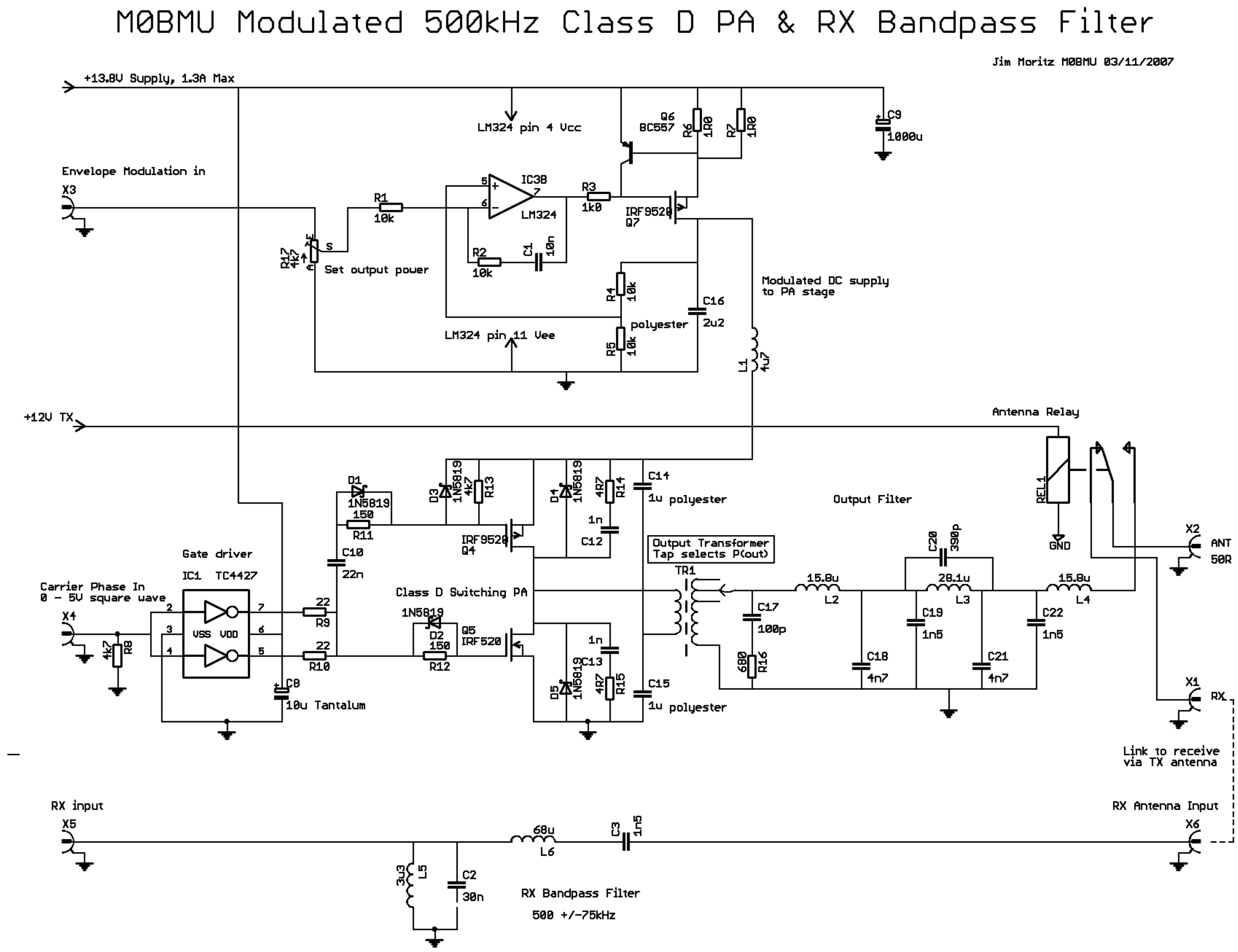


Fig 6 Class D PA, Modulator and Receiver Bandpass Filter

Q7 - mounted on 4^0C/W heatsink
Q4, Q5 - mounted on 20^0C/W heatsinks

L1 – 4u7 choke – rated at least 1.2A DC
L2, L4 – 53 turns 0.4mm wire on Micrometals T68-2 toroid core
L3 – 70 turns 0.25mm wire on Micrometals T68-2 toroid core
L5 – 24 turns 0.4mm wire on Micrometals T68-2 toroid core
L6 – 68u axial leaded choke
D1 – D5 – 1N5819 1A schottky diodes
RL1 – 12V relay with 3A changeover contacts
TR1 – Output transformer on 22mm Fair-rite #43 material toroid core (5943007601) (RS Components 467 – 4217)
Primary – 3 turns 0.8mm wire
Secondary – 15 turns 0.6mm wire total, tapped to suit required output power,
e.g. approximately 10 turns = 5W
13 turns = 8W
15 turns = 11W

C2, 3, 12, 13, 17 – 22 should be low-loss types, e.g. polypropylene, silver mica, polystyrene, etc.

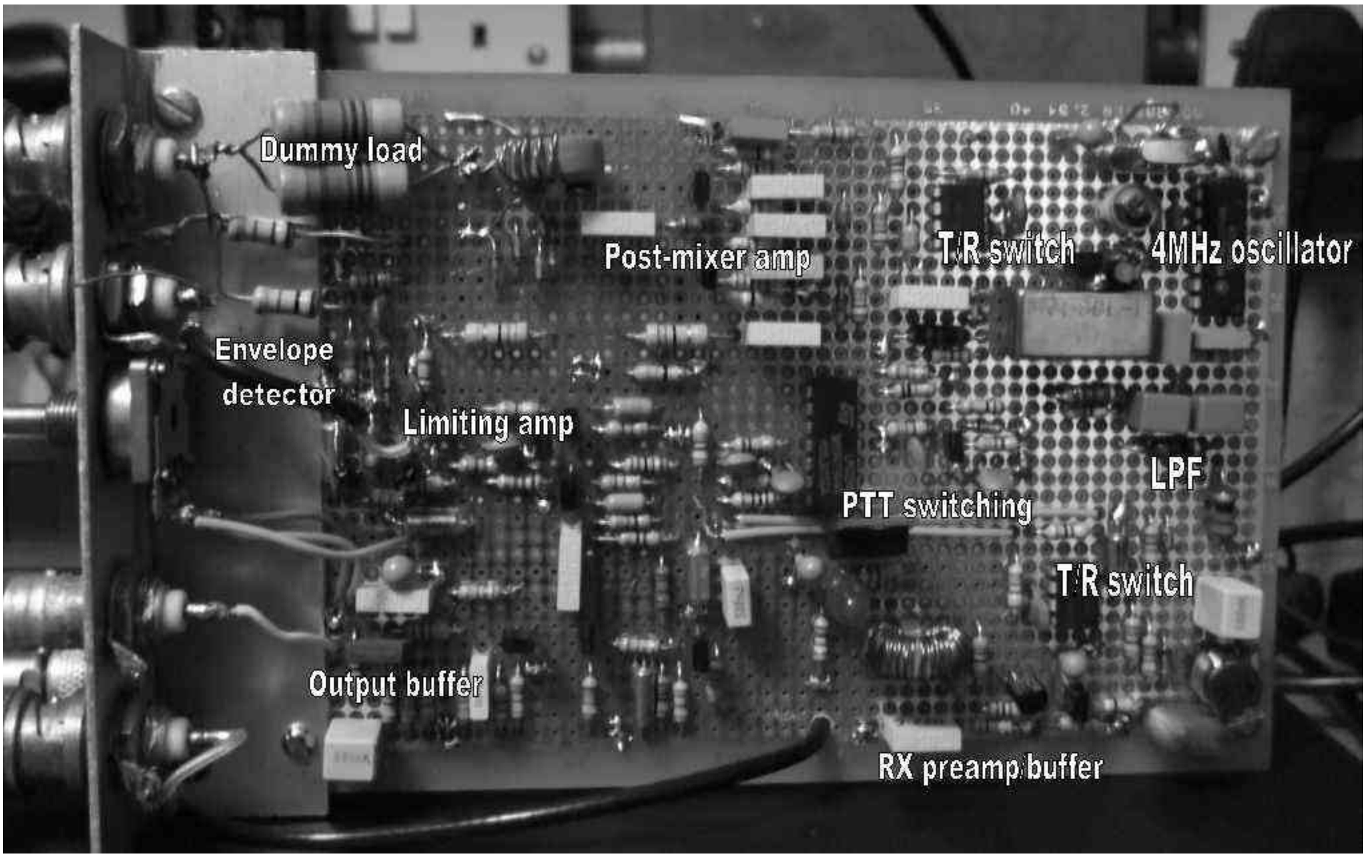


Fig 7 Transverter prototype – built on a "Eurocard" ground plane breadboard (Roth Elektronik RE201) 100 x 160mm. An aluminium bracket attached at the left side carries all the connectors and the linear output level pot. (Note that the schematic shows 2N2222 and 2N2907 transistors – in the prototype, plastic-packaged equivalents (MPS2222, MPS2907) were actually used.)

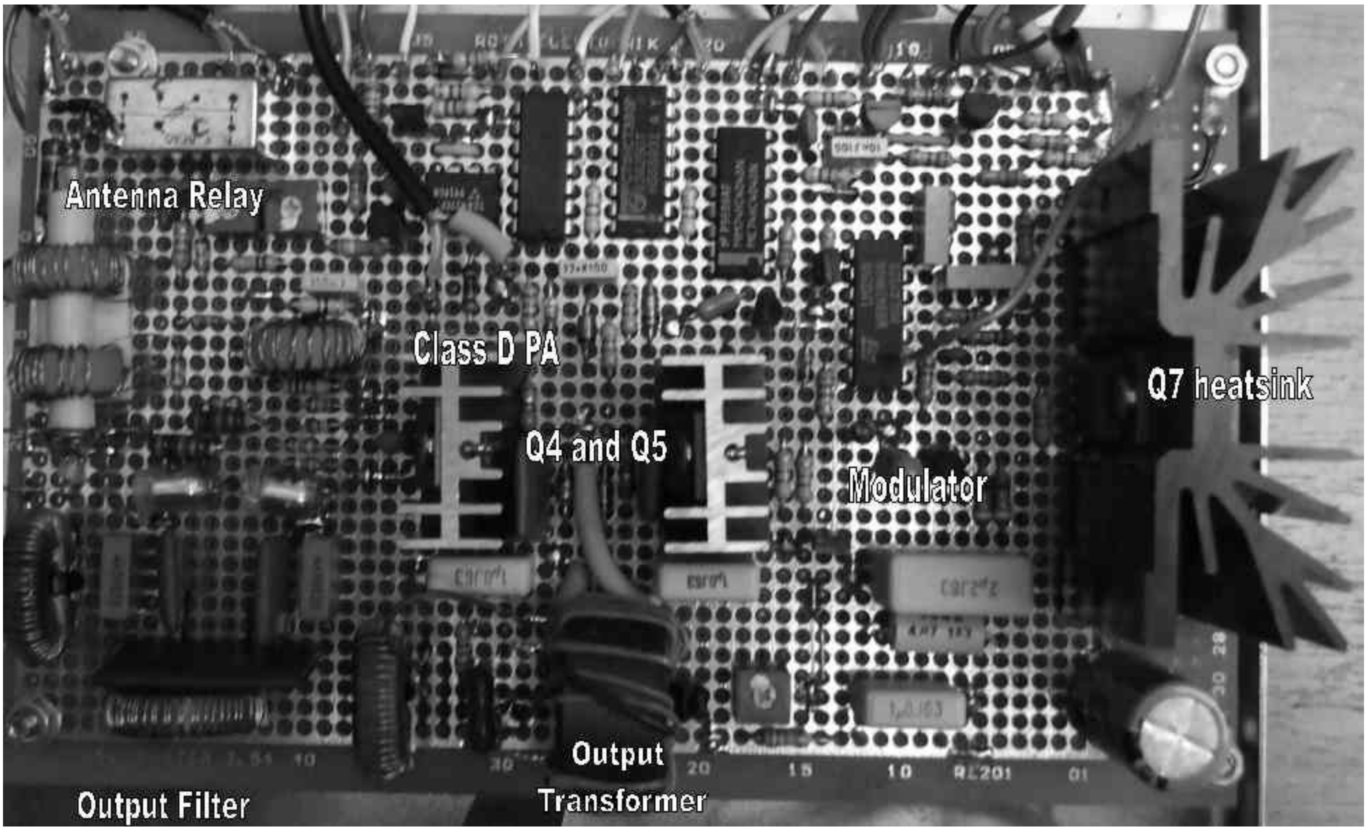


Fig 8 Class D PA and modulator, built on another 100 x 160 mm Eurocard breadboard. This board also contains tuning meter and VFO frequency divider circuits not described in this article.