A 700-W Switch-Mode Transmitter for 137 kHz

This European project makes a lot of VLF power. With our new allocation, it's adaptable for US application, as well.

By Andy Talbot, G4JNT

This project was inspired by the design of the transmitters used for the old Decca Navigator system operating in the 70-150 kHz bands. Decca was decommissioned in 1999, and the hardware from some of the UK-based transmitters became available to the amateur community a couple of years ago. Many of these units were adapted and retuned for use in the 137-kHz and 73-kHz bands, providing RF output of around 1 kW with high reliability under continuous operation for many hours.

The high-power design presented here is not meant purely as a constructional article. Rather, it describes the route I followed to come up with a successful design. It includes enough information and design detail so that experienced constructors can produce a similar unit. *Construction by those without experience in high-power, high-voltage circuitry is not advised.* Particularly, the unit contains some potentially dangerous circuit features such as direct connection to the ac mains and strong RF fields. It has quite a lot in common with modern switch-mode power supplies, however, so anyone confident with these units should have no qualms about constructing this transmitter. Anyone who has built high-power tube amplifiers should be confident enough at these voltage levels.

Decca Transmitter Design

Decca navigation transmitters differ from traditional power amplifiers designed for amateur use in that the Decca units operated at one frequency each, transmitting a pulsed, unmodulated carrier. The constant ampli-

[1]Notes appear on page 26.

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tude of the unmodulated signal meant that a very high-efficiency switching power amplifier could be used. In this type of design, the output devices are switched either fully on or fully off (saturated or cutoff) at the carrier frequency. This means that power losses in the circuit are minimal; the main loss mechanisms are device on-resistance and passive components.

The well-known class-C power amplifier used for FM operation is part way to being a switching design, and can sometimes achieve a dc power-to-RF conversion efficiency as high as 70%. As in all such nonlinear power amplifiers, the very high harmonic content of the generated waveform—a square wave—is removed by filtering. It is here that even higher efficiencies become possible by choosing a filtering system that returns harmonic energy, rectified and filtered, to the power amplifier to be used again. By optimizing the switching topology so that the devices switch at the optimum point in the conduction cycle, the zero-crossing point, efficiency can be improved to beyond 90%. Various measurements made on samples of the surplus Decca units showed efficiencies around 90-95%. In fact, one user even tried to claim the impossible value of 102%—so we can see that measurement accuracy has a lot to answer for!

The basic concept for the Decca transmitters is given in Fig 1. Three identical modules, each delivering up to 400 W, are combined in an output transformer that effectively connects all three in series for an output of 1200 W total. Each module contains four power MOSFETs in a full-bridge configuration, where diagonal pairs of devices are alternately switched on and off. The result is to alternately switch the polarity of the 50-V supply across the load, giving a 100-V pk-pk square wave. The FETs in each arm of the bridge are driven via a very simple transformer, wound on a ferrite pot core with multiple secondary windings, one for each gate. Direct gate drive with separate secondary windings gives the necessary voltage isolation for driving top and bottom devices and makes design of the driver circuitry straightforward.

The output is filtered by a single tank circuit, forming a series-resonant tuned circuit. An inductor-capacitor combination resonated at the desired frequency is placed in series with the connection from the bridge output terminal to the load, resulting in three tanks in series with each of the combining transformer primaries. Because the circuit is series-resonant, only energy at the fundamental frequency can pass through to the load, and the current at harmonic frequencies is blocked. The effect is to cause the power-amplifier devices to switch at the zero-crossing point of the ac waveform, at current minimum, so reducing device dissipation. For optimum filtering of harmonics, the tank-circuit Q—the ratio of the reactance of C or L at resonance to series load resistance—must be as high as possible. However, the voltage across each element of the tuned circuit is magnified by the Q factor, meaning that some quite high voltages can easily appear. Information gleaned from members of the original Decca manufacturing team and examination of the units revealed that a loaded Q in the region of 5 to 6 was used.

If the switching devices were perfect, the design could be as simple as that shown in Fig 1, but the practicalities of MOSFETs require a more complex circuit configuration to avoid blowing up the transmitter. While power MOSFETs can be switched on very quickly after gate drive is suddenly applied, they do not

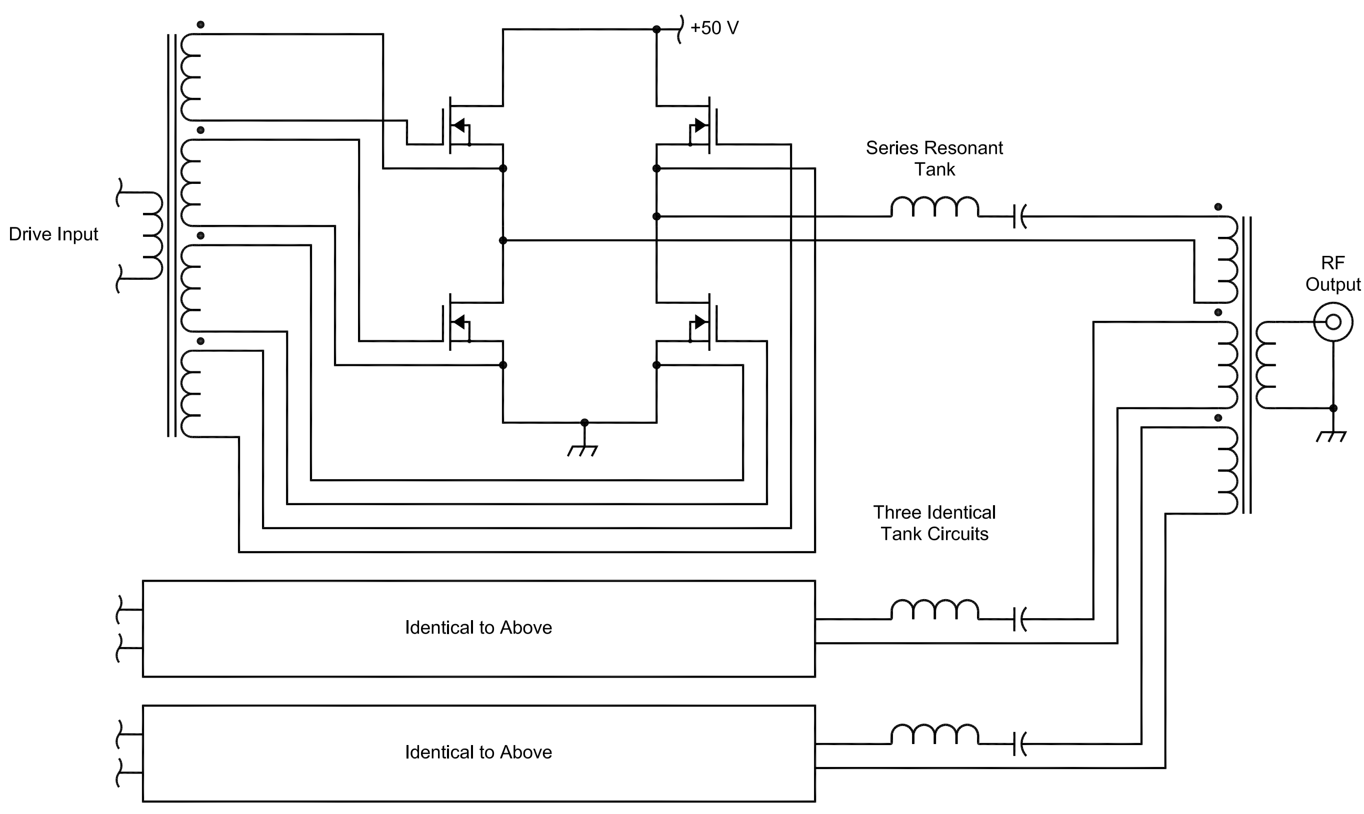


Fig 1—Decca Transmitter outline design.

switch off immediately when it is removed: There is a delay of a few nanoseconds. While this may not seem very long, it results in both pairs of devices being switched on for a few nanoseconds, shorting out the supply and leading quickly to device destruction. In switch-mode power supplies, this problem is overcome by allowing a period of dead time when both devices are off. This period is usually part of the voltage-regulation process in switchers anyway.

Allowing a dead period for the transmitter adds considerably to circuit complexity. The upper and lower devices now need separate drive waveforms as they are no longer both switched alternately, and another approach was adopted here. A small inductor is added between the upper and lower switching devices, as shown in the circuitry around the switching devices in Fig 2. Now, during the short time when both devices are on, the switching transient current merely causes a gradual buildup of stored energy in this inductor that is safely returned to the circuit when the transition is completed. A low-value damping resistor across this kills any high-voltage spikes that may appear should both devices be switched off simultaneously.

The final extra components are the diodes cross-connected from the ends of this inductor to the two power supply rails. The purpose of these is twofold: They return unused harmonic energy to the supply, contributing to the high efficiency of the power amplifier, and they clamp the maximum voltage across each device so that it cannot exceed that of the supply voltage. There is also a very crafty and elegant guard circuit to protect against output short circuits, but more about that later.

Extremely good reliability of the Decca transmitters was maintained by using 200-V-rated devices on a supply rail of 50-60 V, and by choosing devices with a very generous current rating. The design is vindicated by the Decca team statement that only one transmitter ever failed in service in 20 years of continuous operation! However, I had two main objections to using a surplus Decca unit. The main reasons were that I did not have one and all had been sold. The other criterion was the requirement for a high-current, 50-V power supply.

A 700-W Power Amplifier Design

A few years ago when the 73-kHz band became available, I tried making a transmitter based on switch-mode power-supply unit (SMPSU) practice. I directly rectified the mains to give approximately 340 V, then switched this using a half bridge (a pair of 500-V MOSFETs) into a ferrite-core transformer for isolation and impedance matching. Filtering was performed by a conventional low-pass π-network. As I had never seen MOSFETs directly driven by a transformer without extra dc-restoration components, I instead used a proper high-side/low-side bridged MOSFET driver chip, again following SMPSU practice. The design did indeed work to an extent, but efficiency was only around 80% and I blew several FETs accompanied by loud bangs and flashes. While these devices were cheap, and I had plenty from dismantling old surplus SMPSUs, the driver chips that were destroyed each time a FET blew certainly were not! This project was rapidly shelved, and it remained there for several years.

Being taken by the Decca design, particularly the series-tuned tank concept, I decided to make a version powered directly from rectified mains. A very careful examination of the circuit followed, where I was determined to understand fully the precise functioning of every single component. The use of three identical modules was overkill; I did not need the super high reliability this would give, and having three tank circuits as well would just be silly! A few calculations soon showed that by making use of the 340 V possible by placing a bridge rectifier directly across the mains supply, some quite astronomical power levels could be theoretically achieved with just the one output stage. I already had a suitable PSU module on a PC board from an earlier abandoned 700-W SMPSU project. This was made up of a 10-A bridge rectifier, 1000 µF of supply decoupling and plenty of input-filtering and transient-suppression components. Most of these components came from dismantling old SMPSUs, so the design of mains input filtering was just lifted from these surplus units.

Now for a few back-of-envelope calculations to determine the major component values and their ratings. Initially, we will assume the use of low-cost 500-V MOSFETs in full bridge—such as the IRF840—which can switch 3A comfortably. A 340-V rail in the full-bridge configuration could give a 680-V peak square wave, and at 3 A this is nearly 2 kW! For a first breadboard design, even the thought of this was quite scary and I was sure my LF antenna and loading coil would not survive this sort of power without catching fire or melting.

So, how about a half-bridge design again? Here, the switching voltage is 340 V pk-pk; so again keeping to a current of 3 A, we can get theoretically over 900 W, which sounded rather safer. Refer to Fig 3 for the schematic for the complete transmitter.

Tank and Output-Circuit Design

Changing the supply voltage and going from full-bridge to half-bridge topology means the load impedance for the required output power changes from that in the Decca design, so the tank component values will be different. The load impedance is calculated as follows.

Peak-to-peak square-wave voltage across the load resistance, after allowing for voltage droop and losses in the PSU, will end up at around 320 V. The fundamental frequency component of a square wave has a peak amplitude

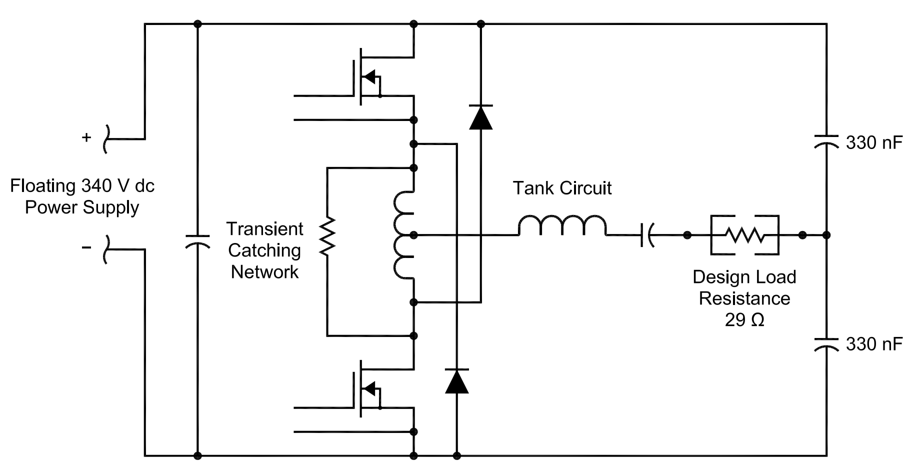


Fig 2—Circuitry around the switching devices.

greater than the peak square-wave voltage by a factor of 4/ð—ie, 1.27 times higher—because of all those odd harmonics combining to flatten the waveform. The resultant filtered sine wave across the load resistance therefore becomes 407 V pk-pk, or 144 V RMS. Having roughly estimated a potential maximum power of 900 W based on device switching capabilities, we need to reduce it a fair bit to allow for rectifier and PSU losses, so assume a power of 700 W maximum. This corresponds to an RMS load current of 700 W / 144 V = 4.86 A—again greater than that of the input square wave because of the 4/ð factor. To achieve this value, a load resistance of 144 / 4.86 = 29.6 Ω would be required, which would be matched to the antenna impedance by a transformer.

For a Q in the region of six, the reactance of the capacitors and inductor making up the tank would need to be about 160 Ω each. At 137 kHz, this means around 7260 pF and 186 µH, respectively, and the voltage across each would be (144)(6) = 864 V RMS, or over 1.2 kV peak. This total capacitance was made up from a series-parallel combination of 1700-V, 3.3-nF polyester caps, with several 220-pF, 1-kV-rated disc ceramic capacitors added across half of the series legs to fine-tune the combination. The tank inductor was wound using PVC-covered Litz wire (obtained from the same source as the Decca transmitters), which could easily cope with 2 kV between windings. The coil form chosen was a piece of drain pipe approximately 44 mm in diameter; coil dimensions were estimated by applying Rayner's formula for single-layer coils:

$$L(\mu\text{H}) = \frac{(ND)^2}{(460D + 1020G)} \qquad \text{(Eq 1)}$$

Where D = diameter, G = coil length (both in millimeters) and N = number of turns. This suggests that around 200 turns would be needed for a single-layer coil, which was impractically long since the Litz wire was nearly 3 mm in diameter. So the coil was wound in three layers and the number of turns adjusted to get close to resonance with the calculated capacitance. The much shorter length and larger overall diameter of the multilayer coils meant that the total number of turns needed was now around only 120.

At this sort of power level, the output-transformer specification could have proved difficult. The largest SMPSU transformer core commonly available, the ETD49 shape using 3C85 material was tried. For SMPSU use, this core is rated to typically about 400 W, keeping temperature rise within acceptable limits. Here, however, the transformer is carrying a sine wave rather than the more usual switching waveform, so it will operate satisfactorily at significantly higher power levels. To calculate the number of turns needed on the primary, the standard equation used with all cored inductors used with sinusoidal waveforms was employed:

$$V(\text{RMS}) = 4.44\, fnA_eB \qquad \text{(Eq 2)}$$

Where f = frequency in hertz, n = number of turns, A_e = core cross-sectional area in m^2, and B is the maximum permitted magnetic field strength for the ferrite used. With the core specified, A_e is 200 mm^2, and B is kept down to 0.1 tesla maximum—well below saturation, which usually occurs around 0.25 to 0.3 T. So for a full-power primary voltage of 144 V using this core, a minimum of 12 turns are needed; to allow a margin, 15 turns were used. The secondary must be tapped to match a range of impedances, from 50 Ω for testing purposes, up in stages to 150 Ω for my antenna in wet weather. Since the power amplifier wants to see a load of around 30 Ω, the turns ratio needed to be in the range $(50/30)^{1/2}$ to $(150/30)^{1/2}$; that is, in the range of 1.3-2.3. So for 15 turns on the primary, the secondary was tapped at 19, 22, 25, 29 and 33 turns. A ceramic switch originally designed for HF ATU use was employed here to switch taps. Remember this item!

The small coil between the upper and lower devices to absorb switching transients came next on the design program. Looking at the coil on the Decca units, and plugging the measured dimensions into Rayner's formula, the value was estimated as 1 µH, shunted by 27 Ω. Well, I was using a higher supply voltage by a factor of over six times, but with reduced current through the devices so the switching transients would not be so bad—let's try making it three times bigger. As I had a few of them, the damping resistor became two 56-Ω, 2-W carbon devices in parallel; although at this frequency, a wire-wound resistor would have been quite acceptable.

Driver Circuitry

The gate drive to the MOSFETs needs to be a square wave with very fast rise and fall times. It also has to be very near to a 50% duty cycle to ensure equal device dissipation and maximum efficiency. Fortunately, there are plenty of MOSFET driver chips around for just this sort of job, and since a transformer is used to drive the gates, the chip would not be destroyed if (or when!) the FET devices blew. (Most driver chips contain a pair of devices, and it may be possible to use both in push-pull to get more drive capability. This has not been tried and is not needed for driving two FETs, but it may become necessary if four FETs are used in a full bridge.) With a transformer in this position, a capacitor becomes essential to remove the dc component from the 0-15 V output supplied from the driver chip. The drive transformer does not need to carry a lot of power, but as it provides the vital safety isolation barrier, it needs to be properly constructed. The windings need to be of well-insulated wire, so a larger core is needed for the turns and insulation than what would have been required to carry the drive power alone. I used an RM10 pot core made of 3C85 material. It turned out to be just large enough to accept 12 + 12 turns of PTFE insulated wire for the secondary (the safety insulation) and 12 turns of normal enameled wire for the primary. This is all a bit tight and an RM12 size core would be better in this position.

Since I wanted binary phase-shift keying (BPSK) modulation as well as on-off keying for LF use, the MOSFET driver chip was controlled by TTL logic designed to provide four phase states—I had decided to include QPSK as well as BPSK. This necessitated an input drive at four times the output frequency, which came from one of my standard DDS modules,[1] which already had provision for driving at four times the wanted frequency. Any other source—such as VFO, crystal oscillator and so forth—is acceptable provided the drive waveform is TTL-compatible and close to a 1:1 ratio. Another drive circuit making use of a comparator and low-pass filter was tried, allowing use of any arbitrary waveform to drive the transmitter.

A schematic of this alternative drive circuit using a comparator to square up a sinusoidal drive signal is shown in Fig 4. The low-pass filter makes sure that the input to the comparator is a sine wave to force it to generate a symmetrical switching waveform in case a non-ideal drive waveform is applied to the transmitter.

Power Supply

In principle, this need consist only of a bridge rectifier and smoothing capacitors. A 10-A rated bridge rectifier gives ample margin for the

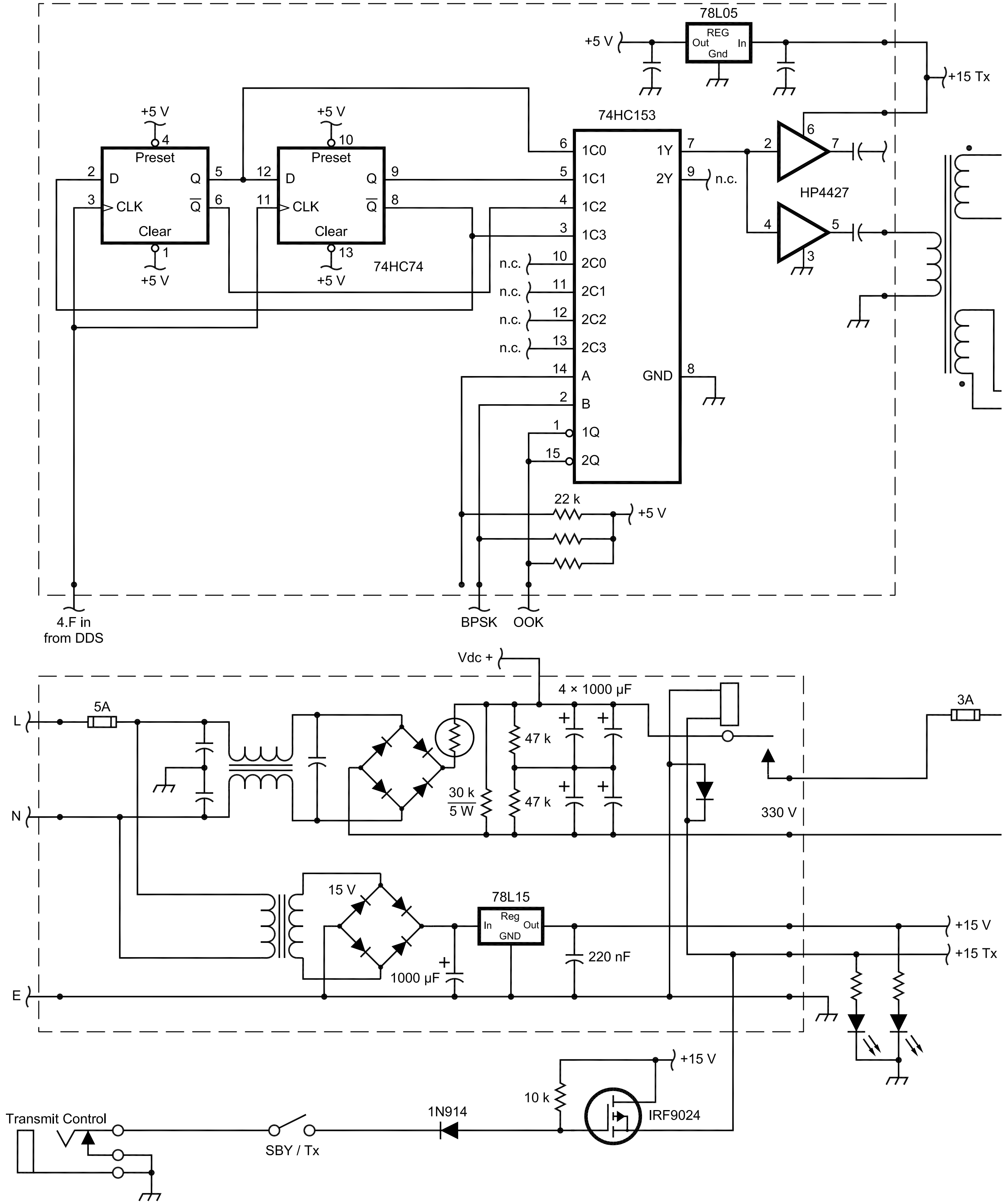
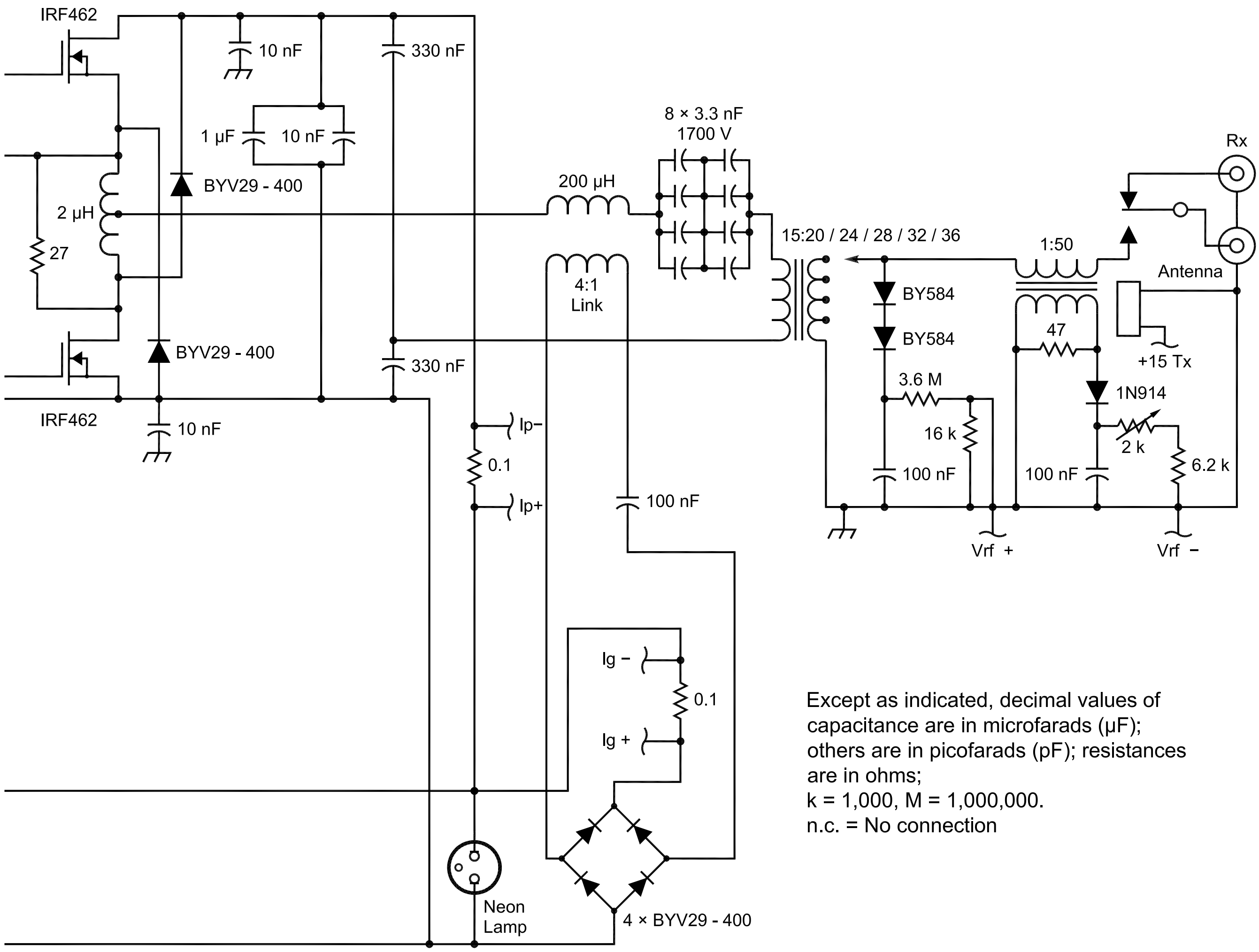


Fig 3—Full 700 W transmitter circuit diagram.



typical 4 A maximum being drawn by this transmitter. The value of smoothing capacitors can be calculated from the equation:

$$CV - It \qquad \text{(Eq 3)}$$

Where C = smoothing capacitor, V = allowed ripple voltage, I = load current and t = the ripple period, 10 ms for a full-wave-rectified 50-Hz supply.

For a switching transmitter, RF output level is directly related to supply voltage, so any ripple will appear as amplitude modulation. Since this transmitter cannot be considered as an AM transmitter under any circumstances, we need to consider the sideband level of the AM components rather than the absolute modulation. A 100% sine-wave amplitude modulation gives sidebands either side of the carrier, separated by the modulation frequency, at a level of 6 dB below the carrier. The value of 1000 µF employed here gives 30 V ripple at a load current of 3 A. This corresponds to approximately 10% ripple, which means 10% AM with sidebands around 30 dB below the carrier. If this is considered too high, the values of smoothing capacitors can be raised. Capacitors rated at 400 V are widely available at values of 1000 µF and higher, but as I had a large surplus stock of 200-V-rated devices, the series-parallel combination shown in Fig 3 was employed.

EMC filtering on the mains input is advisable to prevent LF interference from being fed back along the supply. All switch-mode supplies incorporate such filtering, usually in the form of a dual-wound toroidal choke and mains-rated filter capacitors between the two conductors and from each to ground. The best source of these is often surplus computer power supplies. I also incorporated a thermistor to limit switch-on current. This, too, came from a surplus SMPSU. The last bit of the power supply needed is an isolated low-voltage supply for the driver circuitry and switching relays. Derived from a conventional transformer voltage-regulator assembly, it must supply up to 100 mA for the driver circuitry, plus whatever may be required for fans and relays—typically less than 1 A total.

Safety Notes

As will have become obvious by now, this is a potentially very dangerous project. All the power circuitry is connected directly to the 240-V mains, which, since it is full-wave rectified, means both rectified positive and negative supplies peak at 340 V above ground and average 170 V each. Furthermore, the ac mains supply sometimes has transients on it which can reach kilovolt levels occasionally—albeit just for a few microseconds—because of switching and lightning strikes on some parts of the power network. To enable the FET driver circuitry to be connected to an isolated, ground-referenced source, a mains isolation barrier is essential to ensure there is no direct electrical connection whatsoever between the two circuit halves. This is very conveniently provided by the multi-tapped driver transformer. To ensure proper insulation standards, the windings are wound with good quality PVC- or PTFE-insulated wire rated for mains connections. Similarly, the output circuitry needs isolation. Here, the output transformer performs that function. The primary was wound with the same PVC covered Litz wire as used for the tank coil, which has a voltage rating far in excess of what is required for mains safety isolation.

The final safety issue is that of grounding. With such a large part of the circuitry being connected direct to mains, all metalwork surrounding the finished unit should be very firmly bonded to the mains earth. If a rack-mounted type of enclosure is used, check that each individual metal panel making up the mount is properly bonded. The electrical connection between the metal components is often poor because of their slide fit into anodized aluminum channels and the use of plastic captive nuts.

For all testing, use an isolation transformer! If you need to use a scope on the power-amplifier circuitry, it is essential.

First Tests

At this point, I was satisfied the design was sound and made the first lash-up breadboard on the workbench (Fig 5). The FETs used were surplus IRF840 devices, of which I had many—this fact was to prove extremely useful at this stage! I used a 3 A, 50-V PSU instead of the 340-V rectified mains supply, so I could check out the switching waveforms with a scope and ensure all the circuitry appeared to work as it should. The output load consisted of an old Navy dummy load made up of twelve carbon resistors and (allegedly) rated for 1-kW dissipation. With this 50-V supply, the power amplifier duly delivered the 18 W that would be expected from this voltage rail into the design load impedance, so I was satisfied all was correct. The next stage was the full voltage test. A 1:1 isolation transformer was used to allow direct scope measurements of all waveforms. This was followed by a Variac to allow the supply voltage to be slowly wound up to maximum. Drive was applied, the dummy load connected and (with some trepidation) the supply was slowly wound up while the output voltage waveform, supply voltage and current were continuously monitored. At 100% on the Variac, 320 V was measured was the supply and a sine wave in excess of 230 V peak was across the dummy load—it was working! After several minutes, it was still working and the dummy load was getting quite hot, but so was the heat sink on which the two switching FETs were mounted. Then suddenly, a loud bang and flash and the input fuse blew: Both FETs had shorted. These were replaced, the whole lot tested again by winding up the supply voltage slowly and all worked as before, until—you've guessed—another flash and bang.

It was obvious that the IRF840 devices were being overrun. I also noticed that when probing around the bridge connections with a scope probe, there were a few high-voltage transient pulses at the switching time. Perhaps insufficient decoupling was the problem? A few 10-nF, 1-kV ceramic capacitors were connected across the supply rails close to the FET connections. I also added a couple between $+V_e$ and $-V_e$ rails to the grounded heat sink just in case. Sure enough, the high-voltage transients were killed and by using a fan to cool the heat sink, over 600 W could now be produced for several hours. After this prolonged testing at full power, the output transformer was staying comfortably within its working temperature, so the choice of core and windings was justified, even though it was theoretically working significantly above its specified power rating. I was getting more confident

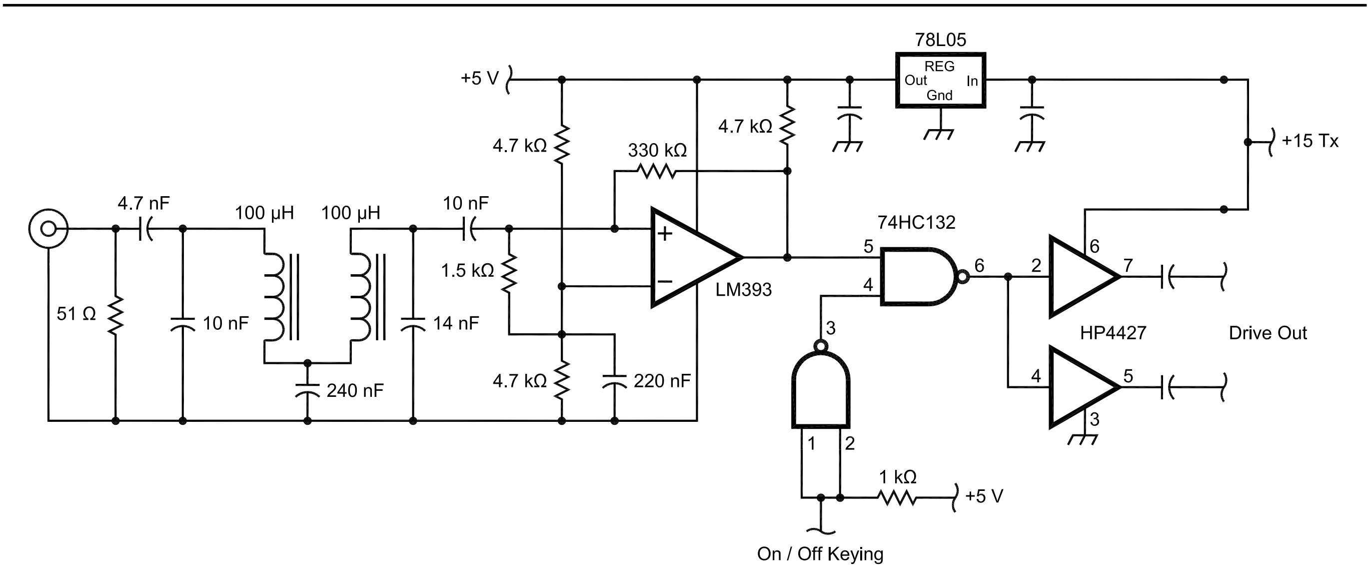


Fig 4—Alternative drive circuit for external input signal.

that the transmitter may actually work out!

How about changing the transformer taps to increase power output? As I moved the output switch to its next tap setting while the transmitter was running, a loud flash and bang! I had failed to notice that the rather nice ceramic switch designed for ATUs was make-before-break. As the switch position was changed, there was a brief short circuit across the three turns between the tap positions and this momentary overload was more than sufficient to blow the devices.

At this point, morale fell and I was not too convinced that the design was going to be particularly reliable. The FETs could easily be upgraded to solve the heating problems, but if they were going to be blown by even the briefest of output overloads, this was completely unacceptable for a finished design. I considered various ideas for overload protection, such as supply-current trips, but none could really be considered perfect. It was only after mentioning this problem on the LF e-mail reflector that Jim Moritz, MØBMU, replied with "Have you looked at the Decca protection circuit?" It had never occurred to me that the Decca transmitters would have had exactly the same overload problem! No protection circuits were shown on the simplified diagrams of the Decca units I had examined. Jim had the full diagrams, having obtained one of the original units, and sent me the details. He had also worked out how the protection operates.

Overload Protection Circuitry

Refer to Fig 3. At first sight, this is a rather unusual bit of circuitry to see around a transmitter power amplifier. A second winding over the tank coil feeds via a capacitor to a bridge rectifier; the dc output from this feeds back to the supply rails. How can this provide overload protection?

The functioning of it is as follows. As RF output current through the tank coil rises, the resonant voltage across this rises proportionately: Remember that as the tank has a Q of around 6 in normal operation, the voltage is already in the kilovolt range. Now, arrange the turns ratio of the over winding to give a transformation ratio such that at maximum rated load, after full-wave rectification and smoothing, the dc voltage produced is equal to the supply voltage. The capacitor in series with the link winding is there to tune out its reactance, but operates with a very low loaded Q, in the region of one or two, so no adverse resonance effects are seen.

Now, if any attempt is made to draw any RF current through the tank exceeding the maximum design value, the rectified voltage would try to rise above the supply voltage. As this is directly connected to the dc supply, it obviously cannot rise above the nominal 340 V. Instead, the rectified power feeds back into the dc supply. This is where things get interesting. Power is now being taken from a component—an inductor—that ideally would not be dissipating anything. The effect of progressively taking more power from the tank is exactly as if a resistor were to be added in series, whose value increases as the overload goes up. The effect is nearly equivalent to operating the transmitter at a constant output current, equal to the maximum rating. Furthermore, by monitoring the dc current being fed back from the guard-circuit rectifier, the degree of overload can be measured.

Calculation of the link winding is not straightforward. If coupling between the two windings were perfect, that is, as if they formed a transformer, then for a tank-circuit loaded Q of 6 the turns ratio ought to be 1:6. However, for an air-core coil such as this, mutual coupling is never perfect and the only way to test the overload protection is to try various numbers of turns for the link coil and see what works. This was obviously not something to be done when operating at full power, so back to the 50-V, 3-A current-limited supply. I initially estimated (read "guessed") that coupling may be in the region of 40%, so I wound on a secondary winding with a turns ratio of 1:4. Overloading the output now did not cause excessive supply current, and the guard current did indeed rise: It worked. By altering the number of turns on the link, it was soon determined that coupling was a bit higher than I had estimated, at around 60%. It also became apparent that the presence of the extra wire around the main tank coil was detuning it, so the resonating capacitance had to be adjusted by around 10 %. Now, we're ready for the full-power test.

Final Design

Again using the Variac and isolation transformer, power was increased to 600 W, and soak-tested for seven hours: No problems. The carbon resistors in the dummy load were glowing dull red (so much for their 1 kW rating!). The devices on the heat sink were at around 50 °C, the output transformer about the same, and the tank coil was running at around 40°C. Those are all quite reasonable figures, especially during the middle of a UK summer. Shorting the output resulted in the devices surviving and only running slightly hotter than normal; guard current rose and supply current fell as expected. Now, to change the tap setting to increase power output. The power amplifier was supplying 700 W now, above my original estimated design value. The dummy load was getting so hot that I had to put it on a metal plate with a blower.

Fig 5—Transmitter components used for breadboarding: top-left, dc power supply; right-hand side, tank components and output matching / isolation transformer; bottom-left, driver and switching components.

After an hour or so, I tried the next tap. The power amplifier briefly gave around 800 W then blew its output devices. I had finally tested it to its limits and the devices had just gotten too hot to carry on living. So, for reliability with these IRF840 devices, around 600 W should be considered the limit, with 400-500 W for continuous 100% duty-cycle operation. I replaced the IRF840 devices (stocks of these were by now getting low) and this time fired up the transmitter without the mains isolating transformer and Variac.

After another eight hours at 600 W all was still going well, so it was time for an on-air test. After connecting a beacon keyer module to the on-off-keying circuit to send my call sign periodically, the first beacon transmission was started. It operated flawlessly for several hours before I decided that the design was final. It was time to put it into a case to make the finished unit (see Fig 6).

I already had a surplus steel 19-inch rack-mount drawer that would take all the components comfortably. Mounting the tank coil was the biggest problem: It needed to be as far away from the metal case as possible to avoid reducing the *Q* and introducing additional losses. In the end, it was supported, horizontally, by spacers, still a bit near to the top of the unit. Some retuning of the tank was needed as the coil inductance had dropped a few percent from its proximity to metal surroundings.

A small fan was added, blowing directly onto the heat sink and metering was added to measure a number of operating parameters. These included RF output voltage and current into the load, along with heat-sink temperature using a thermistor mounted by the switching devices. (The area is shown in Fig 7.) A resistor bridge circuit was used to give a zero-to-full-scale, almost linear range of 20-70°C. Another prolonged testing session proved the reliability, but I felt the devices were still running a bit too hot at 600 W for a fail-safe design. The intention had always been to replace the 5-A IRF840 devices with higher-current types when the design was proven, and now was the time to substitute more exotic IRF460 FETs. Another prolonged soak test showed that these, indeed, ran a lot cooler even with the output power increased to 700 W. Now, the top of the steel case above the tank coil was getting very hot—much hotter than any other component in the power amplifier!

Remembering how "lossy" magnetic materials can be at high frequencies and that this casing was in the magnetic field from the tank coil, a piece of aluminum was attached to the underside of the steel case over the coil to shield the steel from the field.[2] This was successful in preventing the case from heating up, and the coil Q went up slightly, necessitating a bit of retuning. A couple of turns had to be removed from the guard-circuit winding. The finished transmitter can be seen in the Fig 7.

Operation

After over a year of operation, the work put into reliability and testing has proved worthwhile. There has been no failure of any component after many hours of operation, and I have managed to abuse the unit both deliberately and accidentally many times. I've disconnected it, shorting it, operating with a severe mismatch and once even at the completely wrong drive frequency—400 kHz—by mistake. This latter situation could have had unpleasant consequences as the tank, operating well away from resonance, could easily have allowed voltage or current overload, but fortunately didn't!

Also learned the hard way: Beware of removing the output connectors while the transmitter is operating. Transmitting into an open circuit is

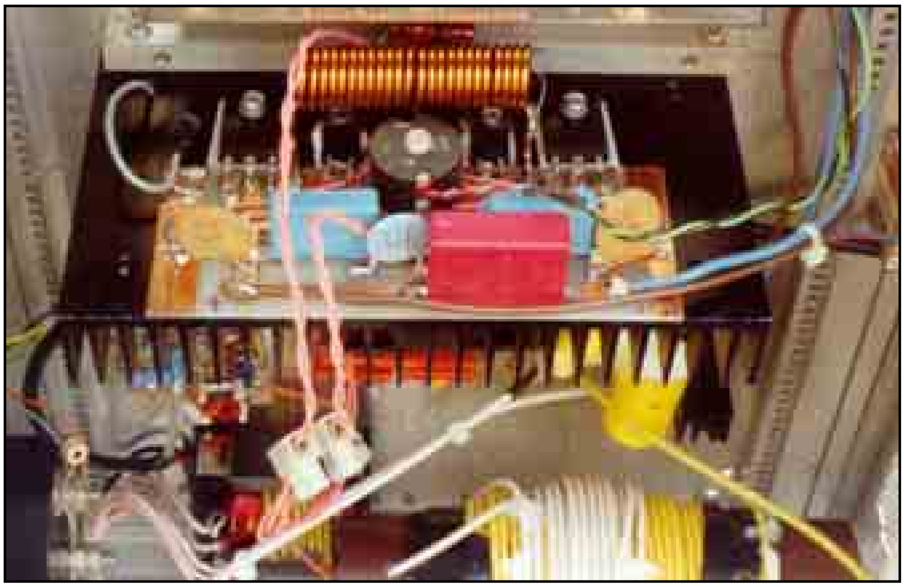


Fig 6—The completed transmitter.

Fig 7—A close-up view of the active switching components.

normally okay, since the transmitter is perfectly happy with a high-impedance load; but once I accidentally removed a BNC connector carrying 700 W of RF. The small arc, created as the inner contact broke connection, triggered a plasma arc between the center pin and the body of the plug, which developed into a sheet of flame spurting out of the plug as the pin vaporized. As I dropped the piece of coax in shock, the arc then proceeded to burn a hole in my floor covering before I was able to kill the power to the transmitter. The BNC plug was a blackened mess with a completely vaporized center pin and insulation. The moral of this story is: Don't under any circumstances remove connectors hot with high-power RF, and use something more substantial than BNC at this power level.

The transmitter has been used with on-off keyed signals such as CW and multitone Hellschreiber, as well as 100% duty-cycle transmissions of binary PSK. All passed through perfectly. In practice, although many parameters are metered, only two need to be monitored consistently during a long period of transmission. Supply current is the main reading to watch, as this is directly related to RF power out. It is immediately obvious when the load match changes. As weather affects the antenna performance, I tend to adjust the output transformer tap position to maintain a figure of around 2.5 A for high-duty-cycle modes, and up to 3 A for short-duration transmissions.

The other thing to watch is heat-sink temperature. At normal room temperature, a relationship between dc load current and heat-sink temperature will become apparent after a few hours of experimentation with varying loads. If the antenna load is significantly reactive, though, the transmitter tends to run hotter than for a purely resistive load. This is caused by the FETs being forced to switch at a point other than zero current, giving rise to increased dissipation in the device on-resistance. Once you have gained practical experience in the supply current-temperature relationship, any discrepancy in this becomes obvious and it is time to check antenna tuning. It may even be worth installing an over-temperature LED or audible indication that trips at, say, 70°C.

Fig 8—Another view of the complete transmitter

Use with US Mains Supplies

The frequency difference of 50 Hz to 60Hz is well-known and just means smoothing capacitors can be 1.2 times smaller for a given ripple at 60 Hz. In the UK and Europe, the as mains is a three-phase supply plus neutral along the street, with 415 V between phases and 240 V from phase to neutral. Other countries are a bit less than this, but rarely below 220 V. Domestic premises are supplied with one phase plus neutral and usually this neutral is connected to ground at many points along the supply route. Full-wave rectifying this gives around 340 V dc; but as one side of the ac feed is grounded, this 340 V is centered on ground, so each supply measures ±170 V mean, but moving at mains frequency. My understanding of US supplies is that a center-tapped 240-V supply is available, giving 120 V for most low-power appliances and 240 V for high-power use. This seems to be borne out by the seven wires I have seen on power-distribution poles.

So, for this transmitter, one option is merely to use the high-voltage, center-tapped supply, in which case the bridge rectification will give easier-to-visualize ±165 V rails that do not oscillate with respect to ground potential. The other option is to use a full-bridge circuit with 120-V input only. Here the transmitter is operated from a 165-V rail, but now four FETs are used in a full bridge as was done in the original Decca design. The driver transformer now has to have four secondary windings, phased to switch diagonally opposite FETs together, and so will almost certainly end up physically larger. The full-bridge configuration has the effect of giving the same peak-to-peak voltage across the load as a half-bridge circuit does off twice the rail. So, load-impedance and tank-component calculations are the same, as are switching currents in each FET. There are just twice as many devices to blow up each time!

Conclusion

The unit described was an attempt to produce a low-cost, easy-to-build high-power transmitter for the 137-kHz band, using surplus components where possible. I was fortunate in having access to several scrap SMPSUs, typical of those used in older PCs, from which many of the power-supply components were recovered. I also had a large stock of IRF840 devices to destroy during the commissioning phases of this transmitter. However, once built and operated within its limits, the design is robust and reliable.

There is plenty of scope within this design for increasing output power. A full bridge running from 340 V will give well over 1 kW and doubling or tripling up the drive units as was done in the Decca design could yield many kilowatts.

Finally, I need to reiterate: *While working on this design, use an isolating transformer right until the end when it is finished and packaged. The voltages and currents can be lethal.* Construct it in a fully enclosed, well-grounded and bonded metal case and maintain good quality insulation and galvanic isolation between the power-switching circuitry and the input-output connections. Also, a few hundred volts of 137 kHz, once it has started arcing, is very hot and creates a strong flame. Plasma-arc welders generate a similar type of waveform as that produced from this transmitter!

Late Note

Since writing this article, the announcement of the US 137-kHz band has been made. As it appears the permitted power is to be limited to 100 W of RF, this design can be used directly from a rectified 120-V supply. With a few changes to the tank-circuit values and output transformer taps using the design guidelines specified, it will supply the maximum power output with ease.

Note

[1]A. Talbot, G4JNT, "A Direct Digital Synthesiser Module for Radio Projects," *RadComm,* Nov 2000.

[2]Aluminum does not really "shield" the steel from the field, but it does intercept the field. The currents set up in the aluminum by the magnetic field tend to cancel that field near the aluminum, and therefore the field strength seen by the steel is much, much less than before. Note that the aluminum must be near but need not be in electrical contact with the steel to do its job.