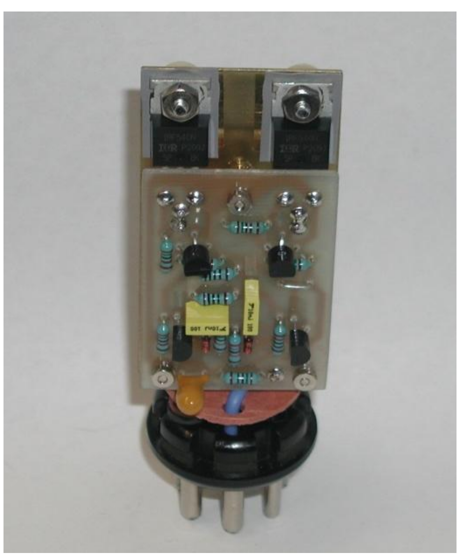
ZC1 MK2: Power supplies. The V6295. Electronic V6295. Tuning frequency metering. Obtaining maximum RF output power into a 50 Ohm load on 40m & 80m. Extension speaker for the ZC1.

Dr. Hugo Holden. June, 2013. (UPDATED June 2014)

The TRANSFORMER ASZ17 OPTION:

THE MOSFET (Osc Driven)OPTION: THE DARLINGTON OPTION: THE SELF OSC MOSFET OPTION:



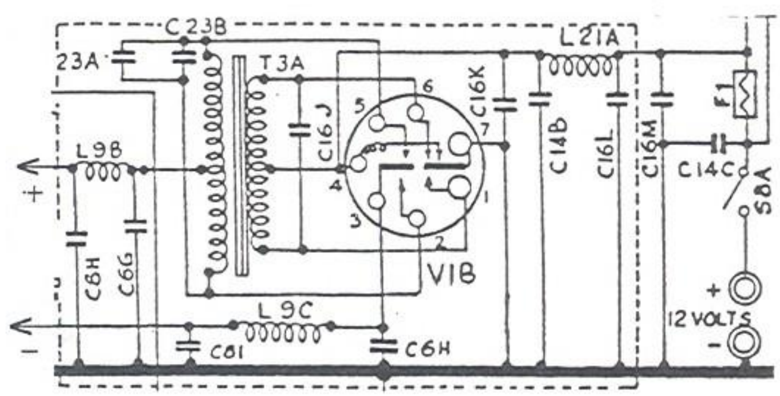
1) INTRODUCTION: THE ZC1 and ORIGINAL V6295 & REPAIRS & ADJUSTMENTS:

For those who are unfamiliar with the New Zealand made Military Communications Radio, The ZC1 MK2, a photo is shown below. For the others, they require no introduction except to say that this radio was a masterpiece of electronics and mechanical engineering. The ZC1 Mk1 was created by the Collier & Beale Company of New Zealand and the MK2 upgraded design attributed to J. Orbell of Radio Ltd. I'm very proud that this extraordinary radio was created in New Zealand, my original home.

Practically every person in New Zealand learning the art of electronics & radio in the post WWII period would have come across this radio because they turned up in great numbers in the 1950's, 60's and 70's in surplus stores throughout New Zealand. They formed a structure on which the radio enthusiast could experiment and modify and at the same time learn about radio reception and radio transmitting. As a result many of these sets were subjected to extreme modifications. It got to a point where unmodified and original units became quite rare. Many of the parts taken from them formed the cores of other electronics projects.

This article focuses primarily on one aspect of the ZC1 MK2, its power supply. The radio was designed to run from a 12V battery source. The tube heaters were place with pairs of 6.3V tubes in series. Interestingly the total tube number in the set added to 11, so one tube required a series resistor for its heater ballast.

The HT supply was provided by an electro-mechanical switching device with a vibrating reed and contacts known as a ***split reed synchronous vibrator***, in this case a 7 pin unit, type V6295. The circuit of this "vibratorpack" power supply is shown below:

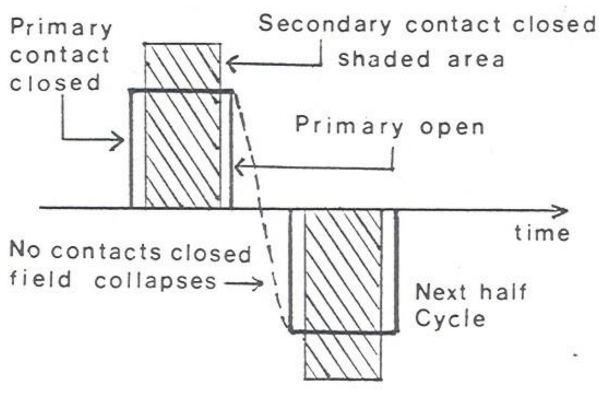


ZC1 MKII POWER PACK.

The V6295 has a pair of contacts to switch the primary winding and another to switch the secondary winding for synchronous full wave rectification. One other contact in the unit is used to switch the magnet coil on and off to sustain mechanical oscillations of the vibrating reed at around 100 cycles per second. This system was quite efficient as the coil in the unit only consumed about 2 watts of power and the contacts, when closed, have a very low electrical resistance. However, in common with all mechanical contacts which switch an inductive load, the contacts would wear and burn and degrade after some 10's to 100 of hours use. There is also another problem related to Latex Rubber inside the unit (see below).

There are a number of articles available on how to repair the V6295. It involves cleaning the contacts of all oxides, having their surfaces mate in perfect opposition when they close and having the correct contact gaps. The adjustable small contact for the vibrating reed is normally adjusted for maximum amplitude oscillations consistent with good starting, however it has a role in very fine adjustment and contact switching symmetry.

If the primary side contact gap is too large the power pack output voltage drops off (as the duty cycle is reduced) and if too narrow the contacts arc over. In addition if the contact gap is too large there is excessive overshoot of voltage on the leading edges of the transformer's primary winding connections. Also the primary contacts must have a slightly longer duty cycle than the secondary contacts and overlap the time that the secondary contacts are closed. There is also a brief time where no contacts are closed and the transformer's field is collapsing. With the value of the transformer's tuning capacitors this is such that the voltage overshoot is as low as possible, thereby minimising the contact arcing and voltage spikes. The timing diagram below indicates these features:



Restoring an original V6295:

A) Mechanical considerations

B) Static contact adjustments

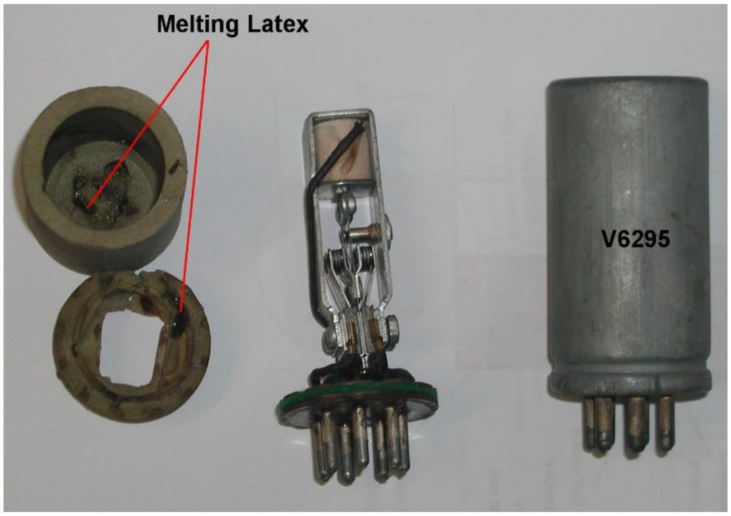
C) Dynamic contact adjustments & oscilloscope recordings.

A) Mechanical Considerations:

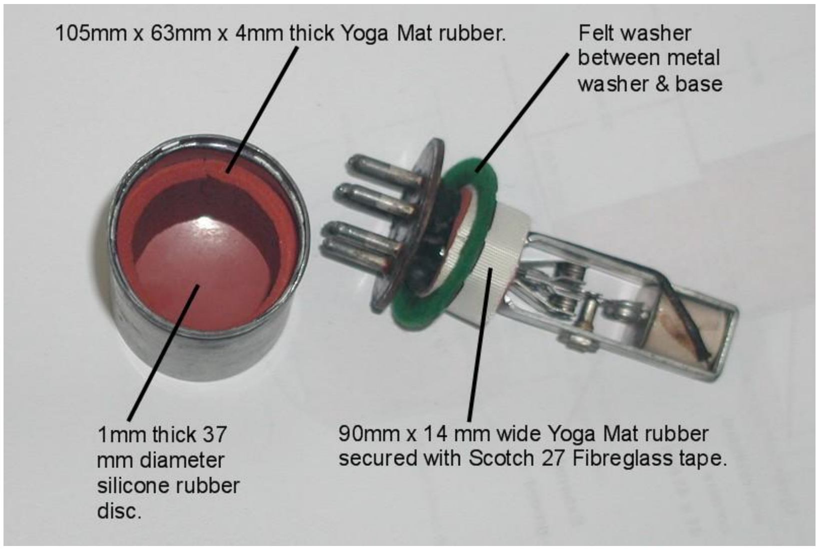
Firstly it is necessary to make an extension plug/socket so as to be able to support the unit while out of its housing and making adjustments and at the same time gain access to the electrical connections:

Secondly the V6295 needs to be in a condition where it can be disassembled and reassembled without damage.

Surprisingly the main reason why V6295's will not run after a period of storage is the latex rubber inside the housing not contact oxidation although the latter is also a factor over longer time frames. As Latex ages and melts with time it turns into a tacky brown liquid and then a vapour. In a closed container such as the metal housing, the liquid goes into equilibrium with the vapour. The vapour deposits out as tacky brown liquid on the contacts as months and years pass. For example an immaculately cleaned and adjusted unit was put into storage. 2 years later it would not run. Taking it out of its housing again, tacky brown deposits had appeared on all the contact surfaces insulating them and causing them to stick together. This material is identical to the areas of melted Latex. Therefore all this old Latex needs to be replaced. The photo below shows a V6295 and some areas of melted Latex. In addition, even without this Latex problem, oxides (which are insulators) reappear on the contacts with time. Also any contact arcing produces very corrosive gasses which are trapped inside the housing.



The rubber inside the unit can be replaced with a number of types of soft rubber product. One thing to note is that the zinc canister was really a little short and there is only a very small amount of room between the mechanism’s top surface and the inside of the zinc case top area. For this top area 1 to 1.5mm silicone rubber sheet is suitable. For the remainder the Rubber from common 4mm thick Yoga soft mat is suitable and easy to get.



When the old latex is removed around the base area it frees up a metal washer which can be separated from the base by a new felt washer (green) shown in the photo. When the unit is assembled back together is important that when held upright, the mechanism can be felt to shake back and forth in the housing. Excessive mechanical coupling of the mechanism to its housing results in mechanical vibrations being coupled to the entire radio and makes the unit noisy.

The best way to remove a V6295 from its housing is by gently prising up the zinc material working around the can very slowly until the lip is unfolded. Then it is carefully smoothed to remove and marks. To re-fit the unit to the housing a brass wire ring can be soldered into position:

This way it can easily be removed later for more repairs/adjustments. **Do not re crimp the zinc can or it can only be cleaned and repaired once as the zinc casing fractures.**

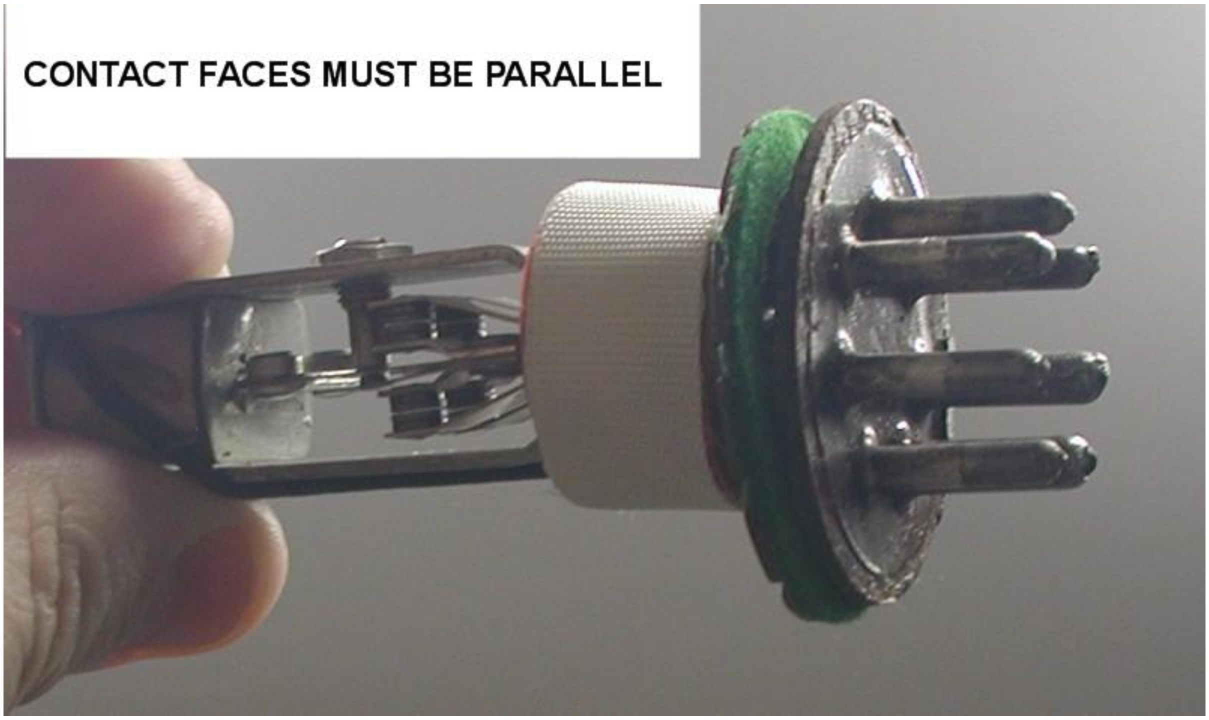
B) STATIC CONTACT ADJUSTMENT:

A first approximation to correct contact settings can be achieved with setting their gaps:

The static gaps for the contacts are 0.1mm - 0.15mm primary and 0.22mm - 0.28 mm secondary. The contacts must first have the Latex deposits, if present, cleaned off with contact cleaner and paper strips passed between the contacts. The contacts can be cleaned with a fresh piece of a 15mm wide 800 grade abrasive paper folded in half with a sharp fold and placed between the contacts. With gentle pressure to close the contacts both faces are cleaned simultaneously. Then a wash with contact cleaner is required to remove any fine debris.

Never file the contacts under any circumstance as this ruins their flat faces and it will not be possible to have a unit with good output and any longevity.

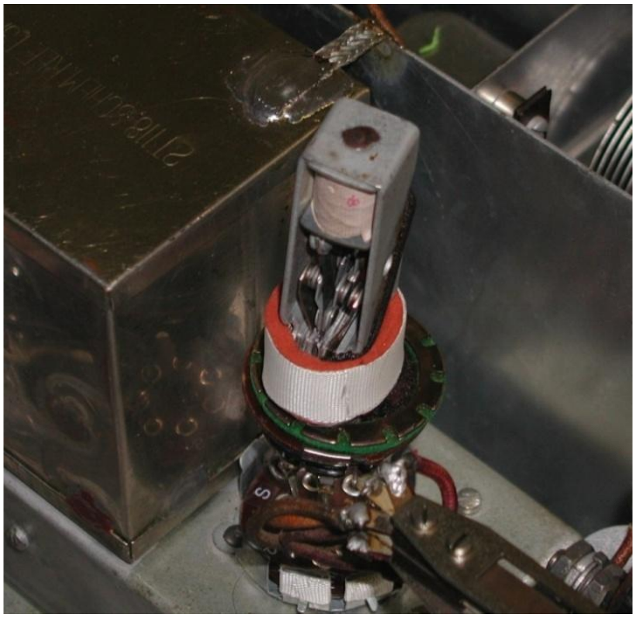
It is important that the mechanical alignment of the contacts is such that when their faces meet their full surface areas are touching and the faces are parallel:



C) DYNAMIC CONTACT ADJUSTMENT:

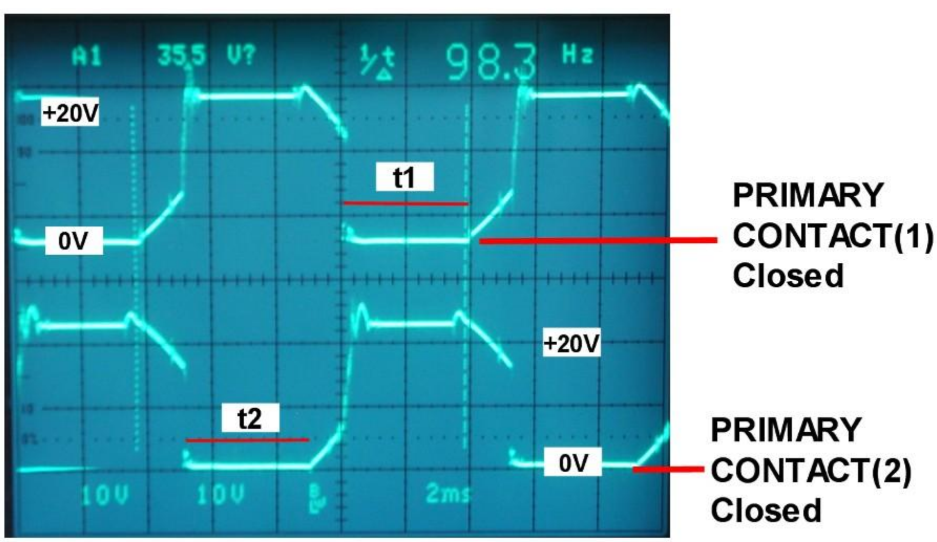
Best output can only be obtained from a V6295 with a dynamic contact adjustment. This involves running the unit out of its housing while monitoring 3 things on the Oscilloscope & meter: The primary connections on the transformer, the DC output voltage and the ripple voltage on L9B.

The photo below shows a V6295 in the extension unit running. The flash photo was taken at a moment when the vibrating arm was deflected. A pair of the secondary contacts were closed at this moment and the other pair wide open, so it makes for an interesting photo:

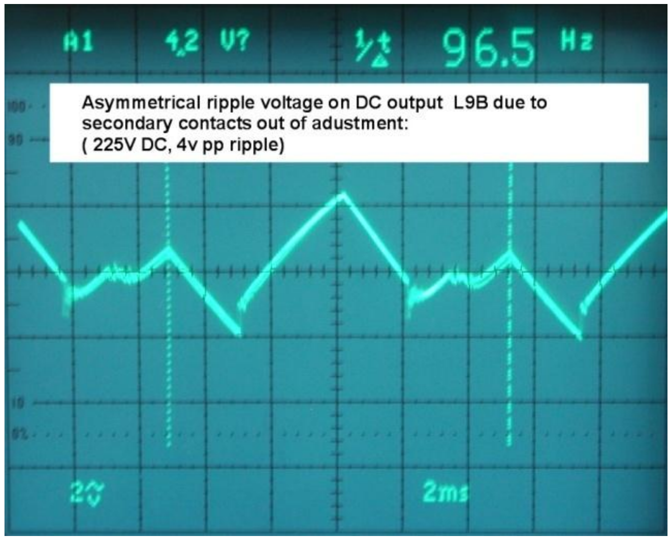


When the ***primary contacts*** are correctly set, each contact is closed for a nearly identical period of time. A small bias can be placed on the contact (with a plastic tool, be mindful of the voltages) to check the effect while viewing the scope. It is important that if the contacts are bent that it is done close to their bases to keep the contacts as parallel as possible. If both the primary contacts are too closely spaced or too widely spaced arcing can be seen between them. If there is asymmetry one will have a small arc and the other not. A *small* adjustment on the vibrating reed contact can correct the centring of the mechanical motion. Larger corrections must be made by moving both contacts. Therefore there is the contact gap to consider and the symmetry of opening and closing comparing one contact to another. The recording below shows the waveforms with correct adjustment of the primary contacts. The time that each contact is closed, t1 & t2, is in the order of 4mS. The photo below shows the scope connected to pins 6 & 1 on the unit. When the primary contacts are closed the voltage on each trace is zero volts:

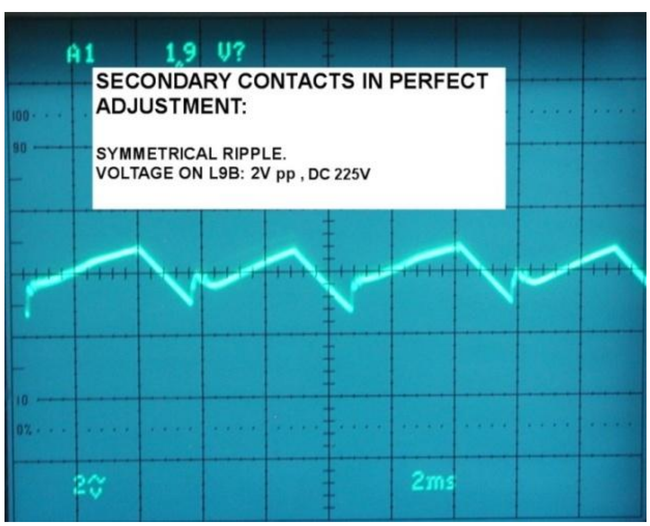
V6295 PRIMARY CONTACT ADJUSTMENT:



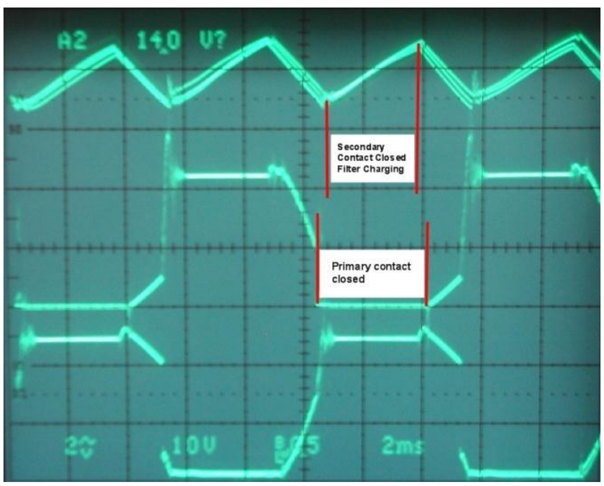
The ***secondary contact*** spacing and symmetry has a profound effect on the ripple voltage superimposed on the DC output (as well as the DC value itself). If the secondary contacts are too closely spaced arcing and flash over occurs. Again the effect on the ripple voltage can bee seen by placing a small bias on the contact with a plastic tool while the unit is running. The photo below shows when the secondary contacts are out of adjustment and how the ripple voltage is very asymmetrical:

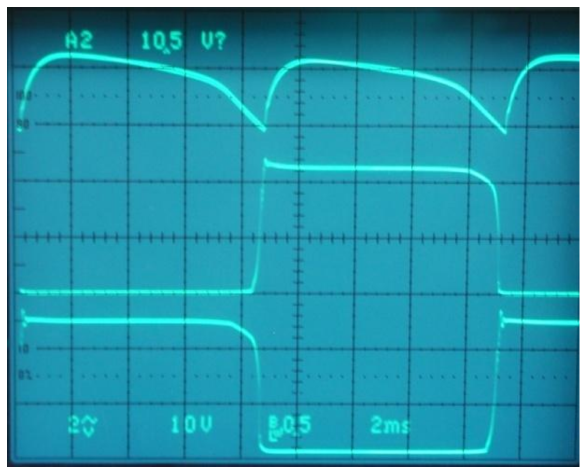


(Note: the DC voltages quoted here are with 11.6V present at the fuse output into the vibratorpack's input and the Sender switched ON (front panel) and the ZC1 in receive mode RT). The photo below shows the correct adjustment of the secondary contacts with a symmetrical ripple voltage.



The photo below (left) is a triple trace, showing both the primary voltages and the ripple output with a well adjusted V6295. Note how the time that the primary contacts are closed is a little longer than the secondary contacts by necessity & due to the contact gaps being wider for the secondary contacts. The multiple traces seen on the ripple voltage is an artefact of the photographic timing versus the scope timing and amplitude variability of the DC voltage. The photo on the right is with an identical test setup and an electronic V6295 pugged in place of the mechanical V6295.This electronic unit runs at 60 Hz (rather than about 100Hz for the mechanical V6295) and makes for an interesting direct comparison. Notice the absence of spikes and transients in the electronic unit and the different shaped ripple voltage which is still about 2Vpp in amplitude:





ORIGINAL V6295 (DC= 225V)

TRANSFORMER/ASZ17 ELECTRONIC V6295 (DC= 235V)

2) THE TRANSFORMER/ASZ17 ELECTRONIC V6295:

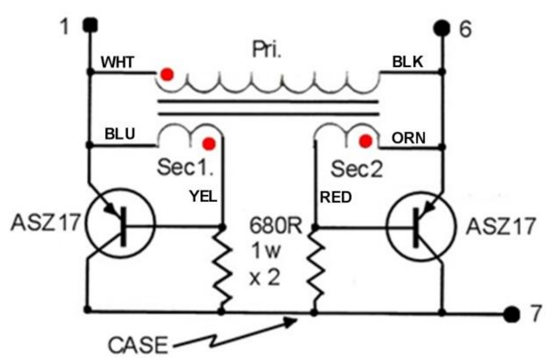
This article shows how the V6295 unit can be replaced by a simple circuit containing two ASZ17 transistors, two resistors and one transformer. Some rectifiers replace the secondary switching contacts.

Power switching circuits which use transistors (BJT's) and do not use driver transformers have large energy losses in the base bias resistors. To give an example of this; the required transistor base-emitter current is in the order of 0.21A. This is because the maximum collector current (the primary side switching current) in this application is in the order of 2.1 Amps (ZC1 in transmit mode) and the transistors must be operated in saturated switching mode. As a rule of thumb, to ensure saturation a 1:10 ratio of base current to collector current is required. If the 0.21A base current is sourced from the fellow transistor's collector circuit, which is transformed up to 24V in use, then the power dissipation is in the order of 5 watts total in the two bias resistors. More efficient transfer of power to the transistor bases involves using a feedback transformer (unless Mosfets are used).

The ASZ17 transistor has a C-E saturation voltage drop of only 0.15v at 2 amps which is favourable compared to its silicon transistor counterpart which has a C-E drop around 0.3V.

The ASZ17/Feedback transformer circuit:

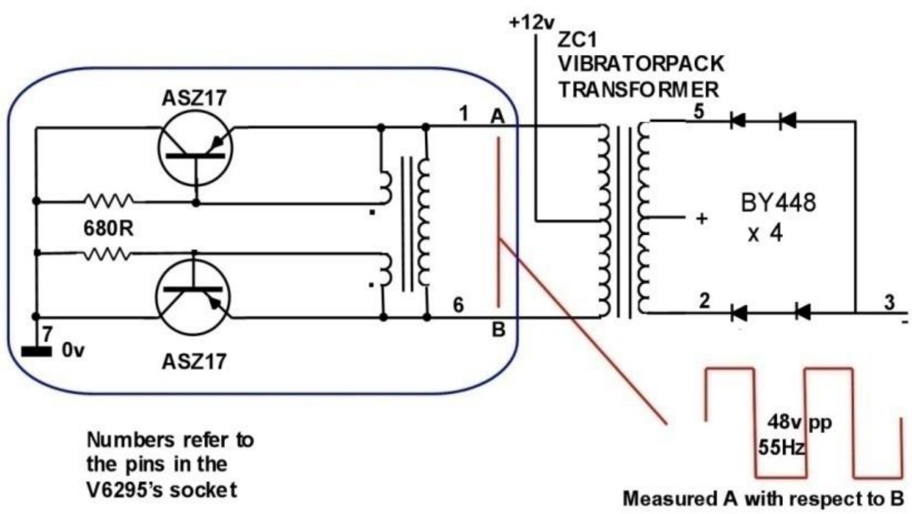
The transformer in the diagram below is a small "feedback transformer" which fits inside a housing which replaces the original V6295. The configuration is a version of the Royer Oscillator:



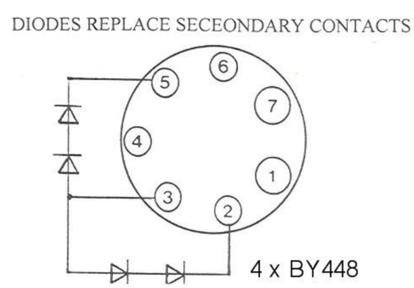
The numbers on the diagram above correspond to the pin number on the V6295 socket. Notice how pin 4 the 12V power supply connection is not used. The circuit is powered by the ZC1 unit's main primary power transformer connections. No DC voltage is applied across this small coupling transformer's primary even in the event of the oscillations stopping with extreme overload. The transformer wire lead colours are also noted here as they match those on the physical transformer.

The feedback transformer transfers the appropriate amount of drive current to each transistor base on consecutive half cycles from a potential which is stepped down from the 24V peak collector voltage to about 3.6V. So the total transistor base circuit drive power, for the two transistors together, is in the order of 800mW. The power loss in the 680R bias resistors is about another 850mW (425mW each). The transistor losses are about 0.3 watts due to their low C-E saturation voltage. The power losses in the four HT rectifiers (in transmit mode output current around 80mA) are about 200mW. So the *total* power loss in this electronic V6295 is only in the order of 2 watts which coincidently is practically identical to the original mechanical V6295.

The diagram below shows the electrical configuration when plugged into the ZC1.



As shown the secondary contacts are replaced by diodes(bottom view of plug):



It is necessary to have a very high *piv* diode rating because if the unit is pugged in and out while running (or a bad connection to one of its socket pins) the undamped collapsing field of the ZC1's main vibrator transformer can produce a peak voltage high enough to reverse break down and destroy a single 1N4007.

A single 1N4007 rectifier is destroyed with this practical test. Two series 1N4007's are required to prevent this. BY448's which are 1500V rectifiers for modern switch-mode power supply applications are ideal, two pairs are required.

How the ASZ17/Transformer circuit works:

Starting from the premise that one ASZ17 transistor is conducting the circuit oscillates because as time passes the ZC1's main power transformer primary side current begins to magnetically saturate the main ZC1 transformer's core, suddenly raising the ASZ17's collector current. The induced voltage is proportional to the rate of change of current with time or dI/dt and this rate of change is falling away with core saturation. Therefore the feedback voltage via the feedback transformer directed to the conducting ASZ17's base circuit drops rapidly along with the base current as magnetic saturation begins. This process is then accelerated via positive feedback and the transistor rapidly comes out of conduction. The drive voltages at the ASZ17's base-emitters reverse polarity and the other ASZ17 transistor is driven hard into saturation. So the process repeats for another half cycle.

On switch on, due to inexact matching of the transistors, the asymmetry in the current encourages initially small sinusoidal oscillations which very rapidly grow to establish stable saturated switching in less than half a second.

It turns out that the running frequency, determined now by the magnetic saturation properties of the ZC1's main transformer core, is about 60 Hz. This is a little slower than the original V6295 which was about 100 Hz. This does not matter, provided the 10uF filter electrolytic capacitors in the ZC1's power supply circuit are new. They can also be replaced with 25uF which is an improvement.

(One point to note about the ASZ17/transformer circuit: The load "appears" to be in the emitters in the circuit diagram of the electronic V6295 above, because the collectors are grounded. However from an electrical perspective the load is actually in the collector circuit & power supply circuit acting in series. This is because the drive voltage is applied by the feedback transformer directly and independently to the transistor's base & emitter connections. This has caused confusion where some think the transistors are

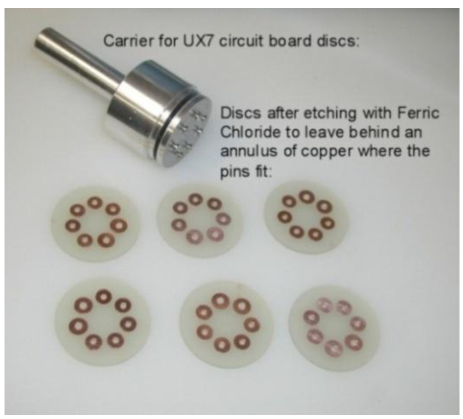
being used in an emitter follower mode and therefore could not act as well saturated switches. An actual emitter follower circuit is unsuited to saturated switching or use in a Royer style of DC:DC converter application)

BUILDING ASZ17/TRANSFORMER ELECTRONIC V6295:

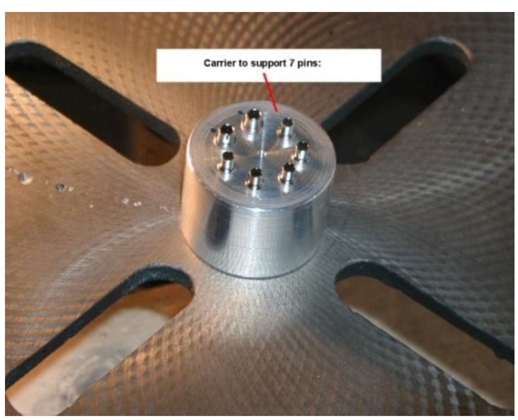
NOTE: Unlike the other electronic V6295 versions presented in this paper, this transformer version is the most challenging for the home constructor to build. The easiest build is the self oscillating mosfet version.

(5 Electronic V6295 units were manufactured at the same time, so there are often 5 items in the photos)

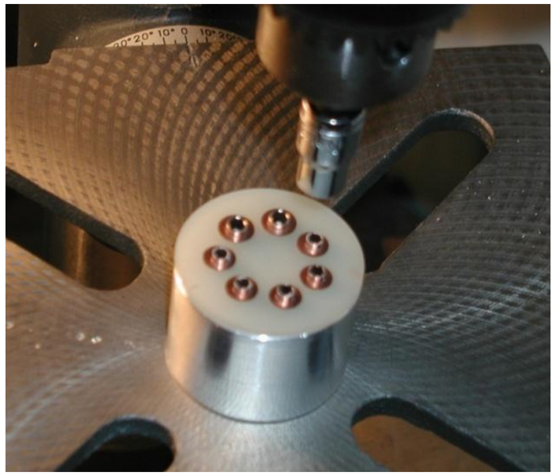
The first step is to manufacture the tools required to make a UX7 base. This is done with two solid aluminium cylinders. The original UX7 base pattern was traced from a scan to make a template to mark the position of the pins. There are two fat pins and 5 thin pins. The tool makes both a carrier for a disc of circuit board material and a template to mark the holes. This can be rotated in the lathe to set its outer diameter at 36mm:



Another aluminium carrier was machined to support the pins while they were pressed into the discs of pcb. The hole for the fat pins was set at 3.95 mm and for the thin pins at 3.1mm. They were then pressed in with the aid of this tool. It was not necessary to rivet the pins in the normal way as the press fit and the soldering to the pcb material imparts the required strength:



The pins are pressed interference fit into the pcb using the drill press and the carrier for the pins and a small socket to do the pressing:



The pins are then soldered to the pcb for additional strength:

The pins were deliberately not riveted as often this can cause splitting if the thin brass material they are made from. The above method creates a very stable and reliable UX7 base and the BY448 diodes are fitted as shown. Only 3 wires pass from the base up into the unit. These are 0.71mm tinned copper with silicone rubber insulation. One is the earth or body connection, the other two simply connect to the ASZ17’s Emitters.

DESIGNING & MANUFACTURING THE FEEDBACK TRANSFORMER:

The transformer for this application requires a specific set of properties. Firstly it must have an iron core due to the low operating frequency. Also it must have a primary winding turns/volt to run a low core flux density because it is important that it is the core saturation properties of the main ZC1 power transformer which determine the operating frequency not the saturation properties of the driver transformer. Therefore during the time of half a square wave cycle (about 8.3 milliseconds) the driver transformer core is well away from magnetic saturation.

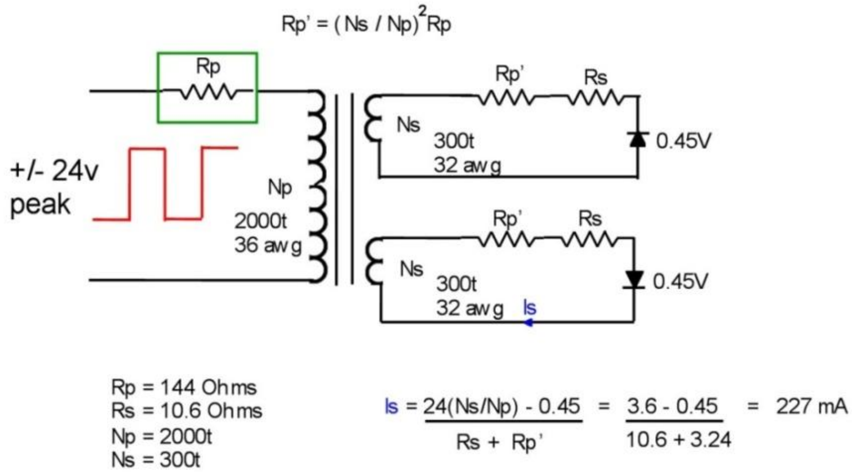
In addition the transformer must have a very specific DC secondary resistance so as to avoid the need to incorporate additional resistors in the transistor's base circuit. Also the transformer must fit inside the machined aluminium housing which replaces the V6295 unit (As shown below this has an internal diameter of 34mm).

The transformer also needs to provide a good base drive current to the ASZ17's base circuits to ensure that the transistors are in saturated switching when they have a 2 amp collector current. This base current is in the order of 150 to 250mA, a typical value being 210mA. A suitable sized core is 1 cm squared inside the bobbin with grain oriented steel laminations.

In this operating mode the feedback transformer's secondaries are effectively shorted out on each half cycle by the transistor's constant base-emitter voltage which is about 0.45 volts. The DC load resistance is of the transformer wire itself. The electrical equivalent circuit for this somewhat unusual arrangement is shown below. For the purposes of calculation, the primary value DC resistance can be reflected into the secondary winding by the impedance ratio (which is the square of the turns ratio).

The primary winding is wound onto the bobbin first, 2000 turns of 36awg wire. Then the secondaries are wound on bifilar. This ensures that they have an identical DC resistance of 10.6 Ohms so that enough DC bias can be developed in conjunction with the 680 Ohm bias resistors to ensure good self starting and also the base current is limited to the correct value as well.

The data below shows how one aspect of the calculation was done. The wire sizes and turns numbers are also such that the full bobbin volume is used with just enough room for the required insulation. The resistances in the diagram below are those of the windings.



The *rms* current in each secondary winding is about 160mA which is over the upper limit for the current carrying capacity of 32awg wire (using the 500 circular mils per amp specification of 126mA for 32awg wire) however the total power dissipated in each winding in this particular interesting case is only in the order of 270 mW and the winding on account of its physical size and external location on the bobbin barely gets warm and there is no threat to the grade 2 enamel wire insulation. The 0.45V forward voltage drop diodes in the diagram below represent the base-emitter junctions of the ASZ17 transistors:

A small lathe with an added turns counter and rpm meter makes for a good small transformer winding machine. With practice it is possible to make the windings very even:

The generally accepted ***flux density*** (Webers per square meter or ***Teslas***) used for iron core low frequency transformers is in the vicinity of one Tesla. The higher this value then the greater the chance of pushing the iron core closer to magnetic saturation. This also depends on the magnetic properties of the iron core, some types saturate before others. As noted it is important in this instance that the feedback transformer does not come anywhere near saturation so it does not play a role in the operating frequency which is determined instead by the ZC1's main power transformer properties. The drive voltage for the feedback transformer becomes, during operation, a 24V square wave @ 60Hz.

The voltage induced in a single loop of wire is proportional to the rate of change of magnetic flux Φ passing through the loop with time t. (The minus sign indicates the induced voltage opposes the applied voltage)

$$V = -d\Phi/dt \qquad \textit{equ 1.}$$

Ignoring the minus sign and separating out the Flux Φ (Units Webers):

$$d\Phi = V.dt \qquad \textit{equ 2.}$$

So the Weber Wb, or magnetic flux Φ, has units of V.s (Volt.seconds). A loop of wire will generate a potential of one volt when the magnetic flux passing through it is changing at a uniform rate of one Weber per second. Or conversely, if one volt is applied to a loop of wire over a time Δt = one second, then $\Delta\Phi$ = the change in magnetic flux inside the loop increases by one Weber.

In the unloaded transformer with an applied AC voltage on its primary, the current lags behind the voltage by nearly 90 degrees. This current has two components, the part which generates the magnetic flux wave which is exactly 90 degrees behind the voltage and a loss component (caused by resistive losses and core hysteresis & eddy currents) which are in phase with the applied voltage. In any event, the flux wave although lagging the voltage has the same frequency and grows from zero to a maximum value Φ_m, in a ¼ of a cycle of the applied AC voltage. This time t = 1/4f where f is the frequency of the applied AC voltage. Therefore the *Average rate of change of flux* is:

Average rate of change of flux = Φ_m/t = $4\Phi_m f$ Webers per second. *equ 3.*

(In a transformer operating with a sine wave voltage a correction is made for the form factor of the wave which is the *rms* value divided by average value which is 1.11 for a sine wave so the 4 in equation 3 becomes 4.44. In this project the transformer is driven by a square wave so the factor of 4 is applicable here)

Webers per second has units of voltage, so the EMF of self inductance which equals the applied voltage V in the transformer with N primary turns is:

$$V = 4\Phi_m fN \quad \text{volts} \qquad\qquad \textit{equ 4.}$$

Rearranging equation 4:

$$\Phi_m = V/4fN \quad \text{Webers.}$$

Since Webers per square meter are Teslas, then dividing both sides by m^2 yields the maximum Tesla value T_m that the transformer core experiences:

$$\Phi_m/\mathrm{m}^2 = T_m = V/4fN\mathrm{m}^2 \qquad\qquad \textit{equ 5.}$$

From equation 5, for the Electronic V6295 feedback transformer, if the primary has N= 2000 turns, V = 24V (square wave), Hz = 60 and for the 10mm square iron core;

$\mathrm{m}^2 = 1 \times 10^{-4}$ square meters:

$$T_m = 24 / 4(2000)(1 \times 10^{-4})60$$

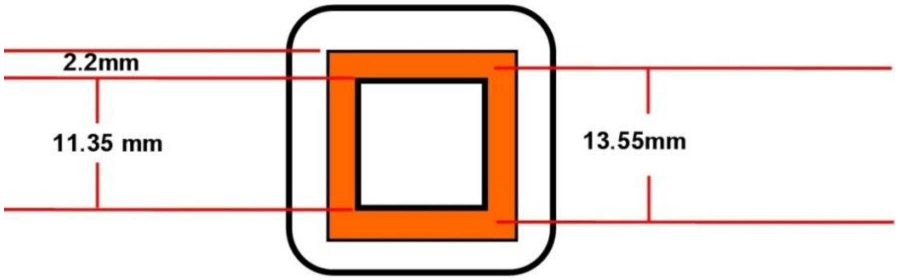
$$T_m = 0.5 \text{ Tesla.}$$

This modest value of 0.5 Tesla ensures that the feedback transformer core for this application is well below saturation.

The transformer's winding DC resistances are possible to estimate from the number of turns and the geometry of the bobbin. The number of turns per layer is closely approximated by the diameter of the wire (including its enamel) divided into the bobbin width. This number is simply divided into the total number of turns to find the number of layers, which is then multiplied again by the wire diameter to calculate the winding height. Once this is known then it is a simple matter to calculate the *average* length of a turn bisecting the centre of the windings and making the assumption that the turn is square(for a square bobbin), then multiplying this by the number of turns to calculate the length of the wire. Then the resistance is found from the copper wire tables.

Giving an example of this for the primary winding:

The width of the bobbin from side to side16.55mm (measured) and the 36awg wire is 0.135mm including its enamel (measured with a micrometer), so there are 16.55/0.135 = 122.6 turns per layer, so a 2000t winding is 16.31 turns high or close to 16.31 x 0.135 = 2.20mm. The inner bobbin, where the winding starts, measures at 11.35 x 11.35 mm. Therefore with a 2.2mm high winding:



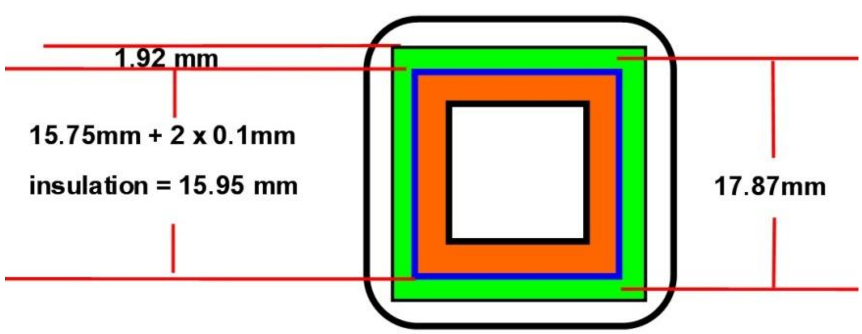
13.55 x 4 = average length of turn
Primary

The average length of a turn is 13.55 x 4 = 54.2 mm and with 2000t the wire length is 108400mm From the wire tables, 36awg wire has a resistance of 2.4105 ft/ohm, which converts to 0.001361 ohms/mm, therefore the expected primary resistance is

108400 x 0.001361 = 147.5 Ohms.

The measured resistance of the actual wound transformer primary is very close at 144 ohms. Using this method of calculation was within 3% of the actual value.

Applying the same principles to the two secondaries which have a total of 600 turns (two bifilar wound 300t windings) the 32awg wire on the micrometer measures 0.23mm diameter, there are 16.55m/0.23mm = 71.95 turns per layer and 600/71.95 = 8.34 layers which is a winding height of 8.34 x 0.23 = 1.92 mm. Adding this on top of 0.1mm insulation tape on top of the primary gives:



17.87 x 4 = average length of turn

Secondary.

The length of the average turn is therefore 17.87 x 4 = 71.48mm and there are 600 turns total, making the wire length 42888mm. From the wire tables 32awg wire has a resistance of 6.0945ft/ohms which converts to 0.0005383 ohms per mm. So the total secondary resistance of the 600t winding is expected to be 0.0005383 x 42888 = 23 ohms.

This makes the calculated DC resistance of one 300t winding 11.5 ohms.

The actual measured value on the wound transformer was 10.6 ohms.

This indicates how closely this simple method estimated the DC resistance, still within 10% of the actual value. The reason the calculations slightly overestimate the DC resistance, more so on the secondary, is that the windings are modelled as rectangular, but in practice the corners are more and more rounded as the winding height increases, shortening the wire length of each turn, more so for the secondary calculation than the primary.

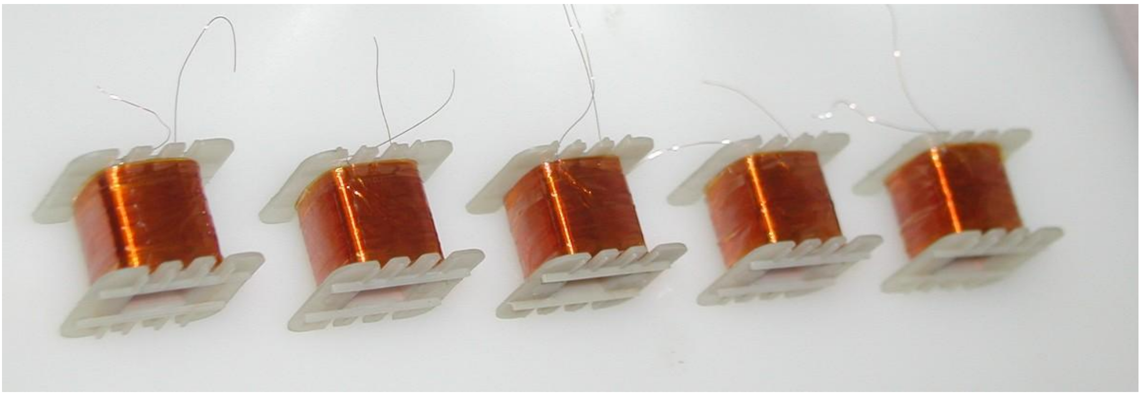
Also from the figure above the total winding height is 2.2 + 0.1 + 1.92 = 4.22mm.The height of the plastic bobbin is about 5.75mm so there is enough room for the outer coat of insulation as seen in the photos.

There are many aspects of transformers not covered in this article, such as leakage reactance & iron losses, other types of ferromagnetic cores etc and RF transformers, winding capacitances and so forth where the rules are a little different.

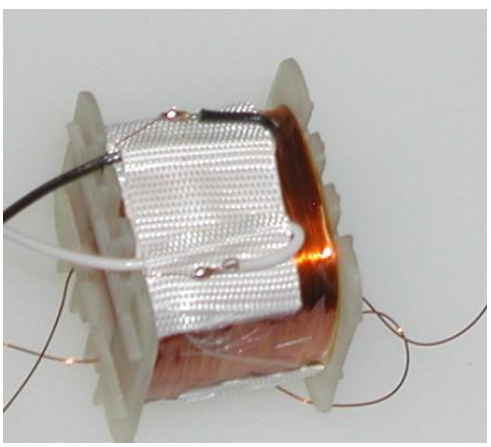
However in general when winding transformers it is important to keep the windings as regular and orderly as possible.

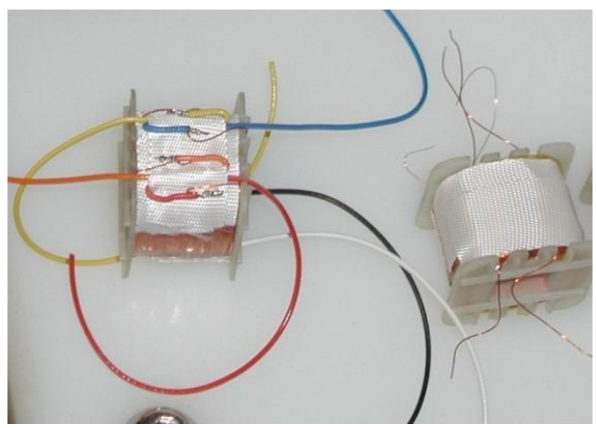
Improved wire enamels combined with factors of economy has meant that the configuration of the typical power transformer winding has changed over the course of a century. Transformers, up until the mid 1960s, even those with very fine wire and 1000’s of turns, were wound in perfect layers, with very thin rice paper like insulation between each perfect layer. There were some disadvantages as residual salts in the paper could, in conjunction with water vapour, cause green spot corrosion of the copper wire. Also they had higher inter-winding capacitances. Still one can’t help but admire the winding perfection seen in these vintage transformers. Such windings are still used in oil filled car ignition coils.

The photo below shows the bobbins for the Electronic V6295 with the 2000 turn primaries wound on. Two layers of polyamide (Kapton) tape are applied:



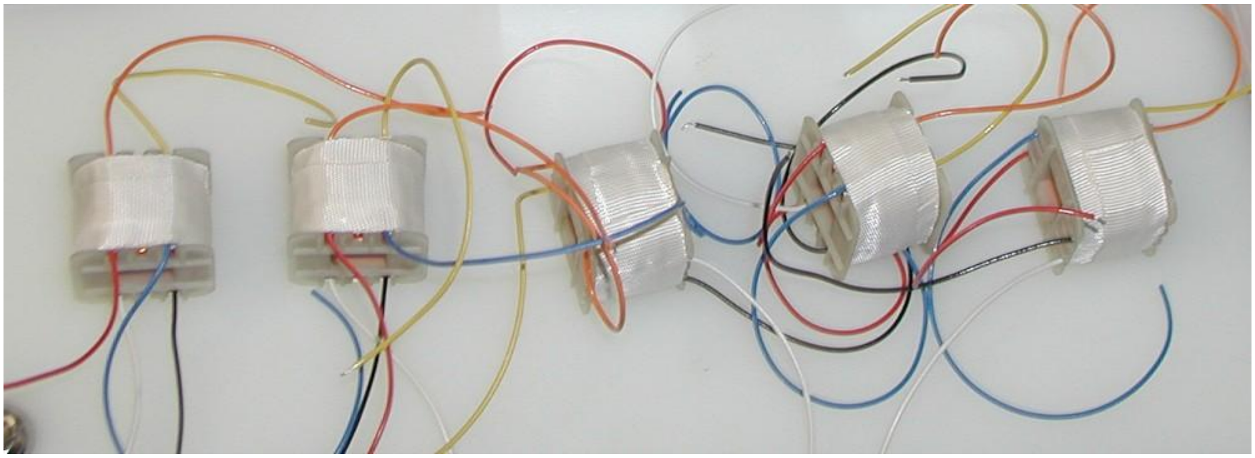
The secondaries are then wound on bifilar and again two layers of Kapton tape. Then some special fibreglass tape (Scotch number 27, made by 3M available from Hayman's Electrical) is used to assist in terminating the wires to their flying leads:



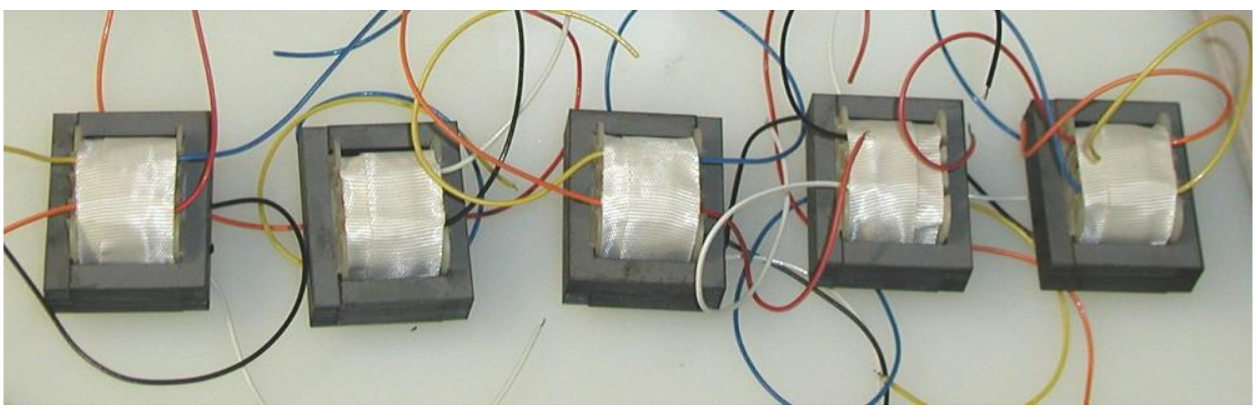


This fibreglass tape is also used to finish the bobbin. It is far superior to the usual yellow plastic transformer tape. The 32awg secondary wire used here has non self fluxing tough grade 2 enamel which needs to be carefully scraped off prior to soldering. The 36awg primary wire has self fluxing enamel.

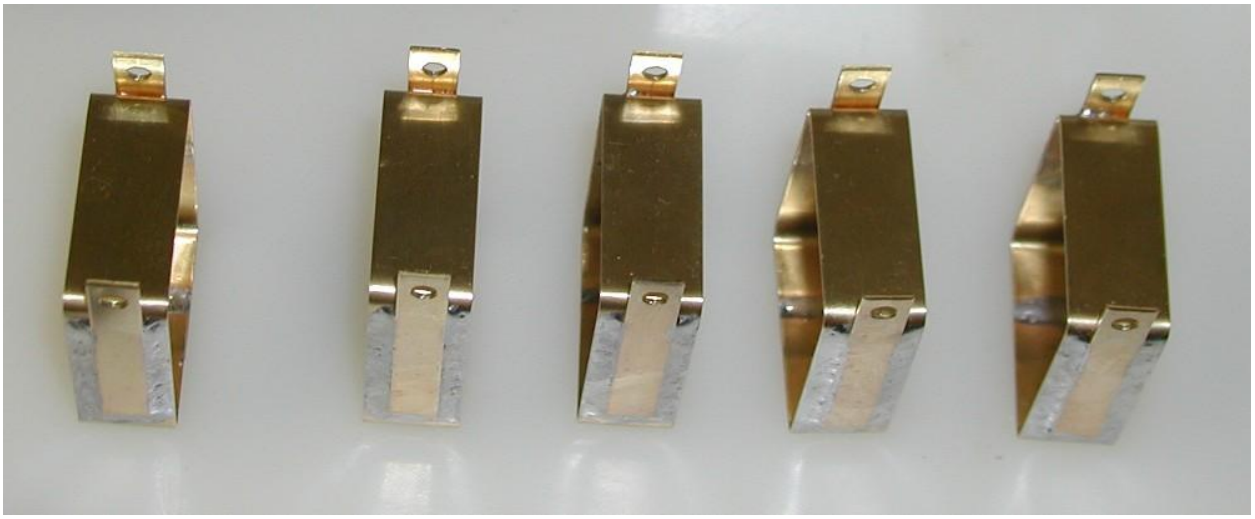
The photo below shows 5 finished bobbins:



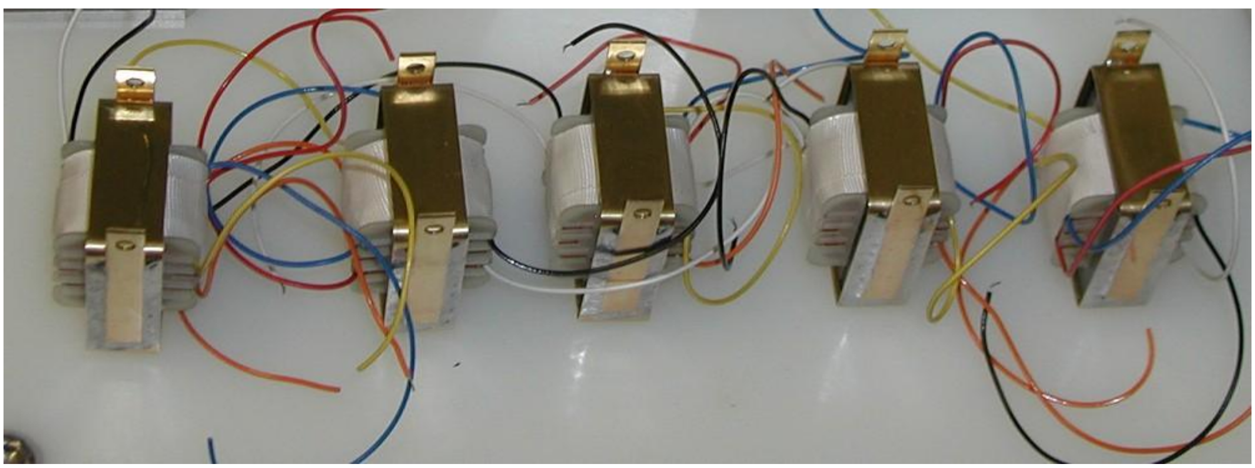
The bobbins are then stacked with their laminations, the edges of which are lightly painted with Fertan organic rust converter. This deactivates any surface rust crystals on the lamination cut edges.



Transformer brackets were prepared to allow these transformers to be mounted inside a 34 mm diameter cylinder. Brass strip ¼ inch wide (0.8mm thick) and ½ inch wide 0.6mm thick is available from K&S Engineering (stocked in model shops). The brass was folded and soldered to create the brackets. The transformer stack is a firm press fit into the bracket and it is also effectively glued to it by the varnishing process:



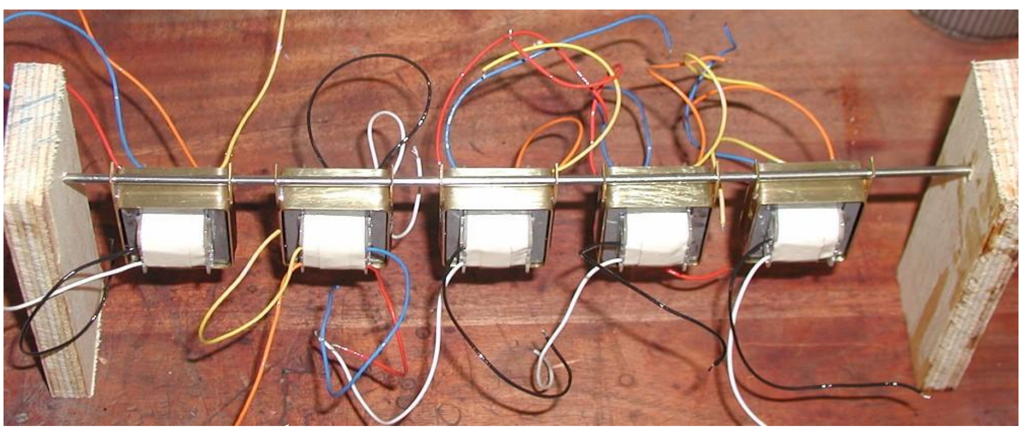
The following photo shows the transformers ready for vacuum varnishing:



While the transformers could simply be dipped in varnish it is better to apply a vacuum. While a full vacuum of 760mmHg is good, it requires a pump. A 500mmHg vacuum can be attained with a simple syringe a strong arm and jam jar as shown below. This causes the air to exit the small spaces in the transformer windings and for the varnish to pass in well. The syringe is pulled upwards expanding a very small air bubble into a large volume. As it is hard to hold it there for long, a brass rod is used to lock the syringe plunger and allow 15 minutes for the multitude of fine air bubbles to exit the transformer:

The photo shows one of the transformers inside the jam jar full of polyurethane varnish and subjected to a vacuum.

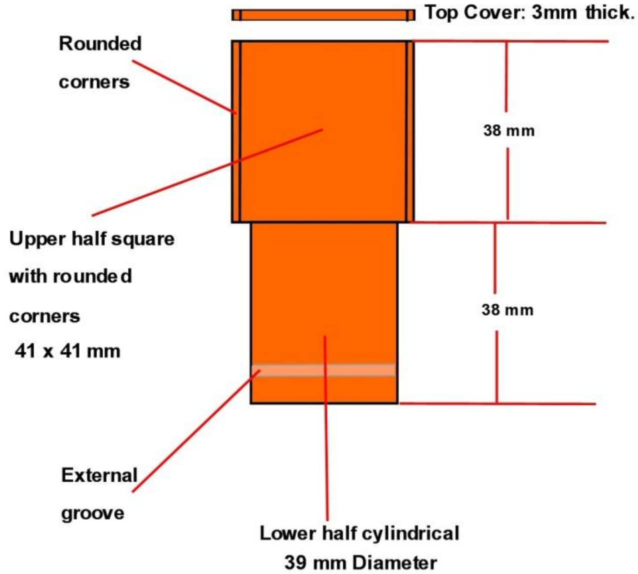
Finally the transformers are hung up to air dry. This process could be sped up with an oven, however I simply left them for one week.

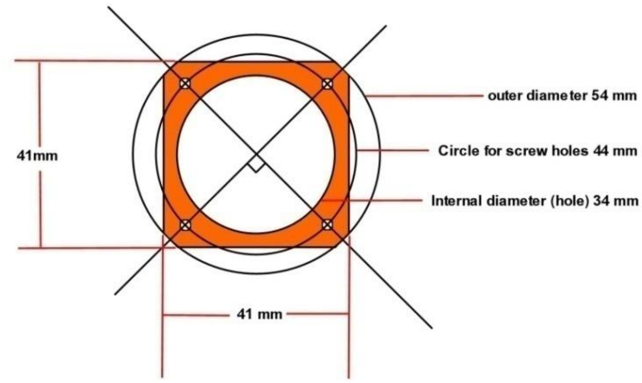


The Aluminium Housings:

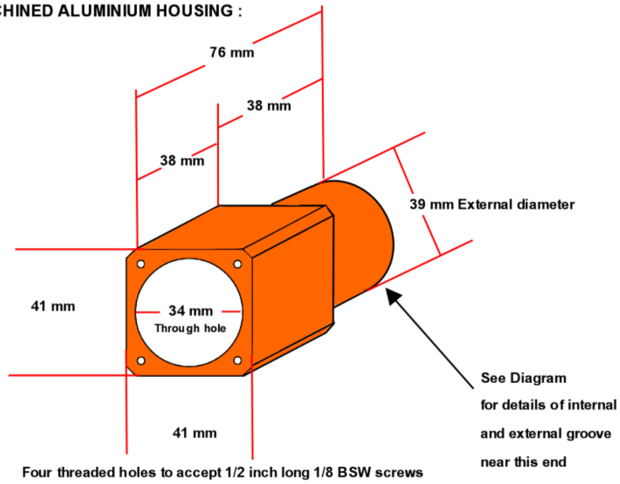
These were made according to the diagrams below, drawn in Picture It, however the manufacturer re-drew them in a CAD program for CNC machining:

SIDE VIEW, ALUMINIUM HOUSING WITH TOP COVER:



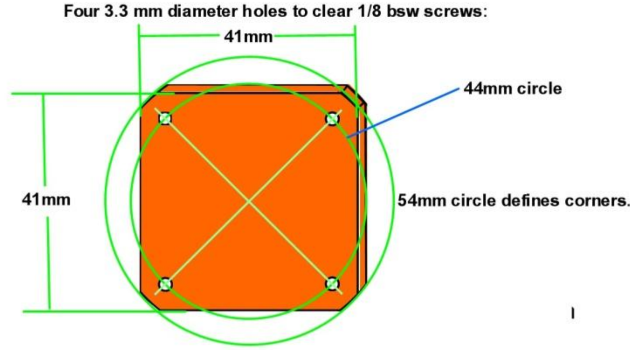


MACHINED ALUMINIUM HOUSING :



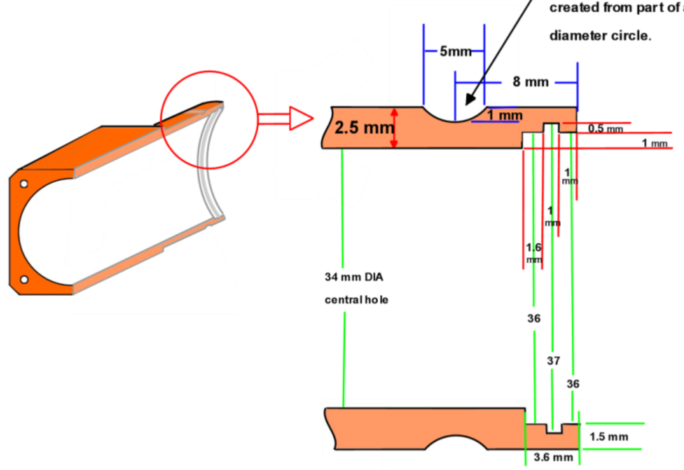
TOP COVER: 3 mm thick Aluminium

Four 3.3 mm diameter holes to clear 1/8 bsw screws:

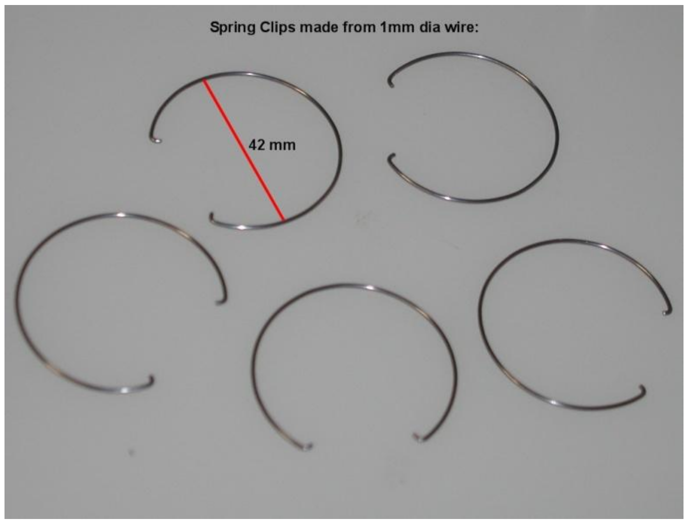


CIRCUMFERENCIAL GROOVE DETAIL :

This depth to width ratio is created from part of a 7.25 mm diameter circle.



The UX7 base is retained in the unit by a wire clip made from #17 Piano string wire which is very close to 1mm diameter and has a springy quality to it. If wound around a 22mm diameter cylinder it springs back to about 42mm and it spring fits into the 0.5mm deep groove in the housing:



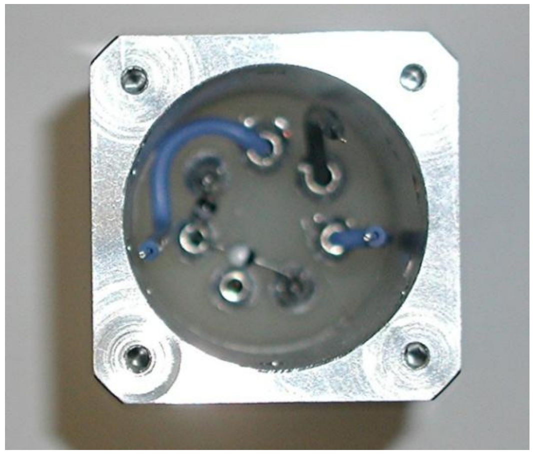
The top cover is fitted with four countersunk 1/8 x ½ BSW screws. The photo shows the housings which were CNC machined for me to very exacting and high quality standards by UP-Machining in Shenzhen China:

The following photo shows the holes added for the TO-3 (ASZ17) transistors. The transistor's mounting holes are threaded with 1/8 BSW. The transformer mounting holes are countersunk. The transistor collectors connect to the case and ground (negative) so there is no requirement for any mounting washers:

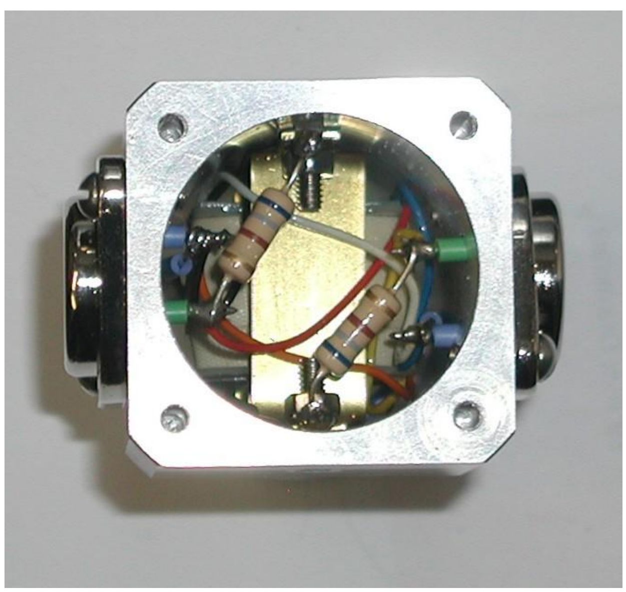
ASSEMBLING THE UNITS:

The 7 pin base is retained in the housing with the spring clip. As it was made a close fit and the based placed into position with polyurethane varnish on its edges and over the clip, it is extremely strong and when the varnish dries impossible to rotate the base in the housing. Although it could be dissolved out one day if ever required. The base is rotated to the correct position before the varnish dries so as to accommodate the rectangular top of the housing when plugged in beside the main transformer in the ZCI.

The photo below is looking down into the unit before the transformer is slipped into place. Only three wires rise out of the base:



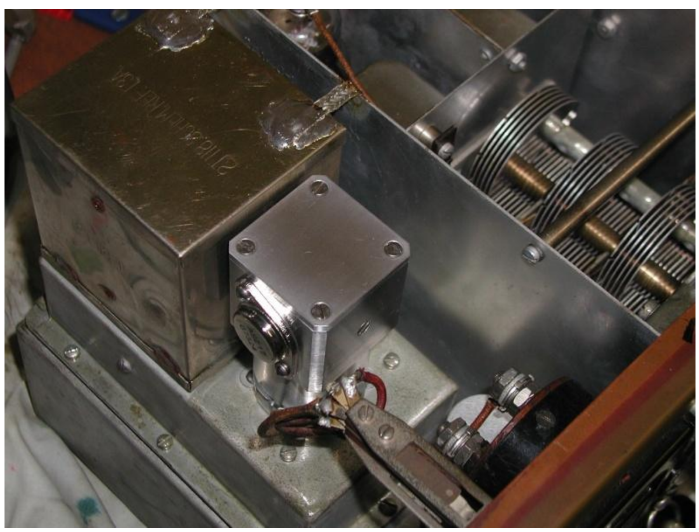
The transformer is placed into the housing and retained by two 1/8 x ½ bsw slot countersunk screws with nuts and spring washers. Earth lugs are placed between the transformer mounts and the inside of the aluminium housing and are the solder tie points for the two 680R 1W resistors and ground(negative) and the black wire from pin 7 on the base. The transistors are fitted and screwed to the case with 1/8 x 3/8 bsw roundhead slot screws. The transistor's base & emitter leads have a protective silicone rubber insulating sleeve, green for the base and blue for the emitters. The emitters connect to the blue wires leading to pin1 and 6 in the base:



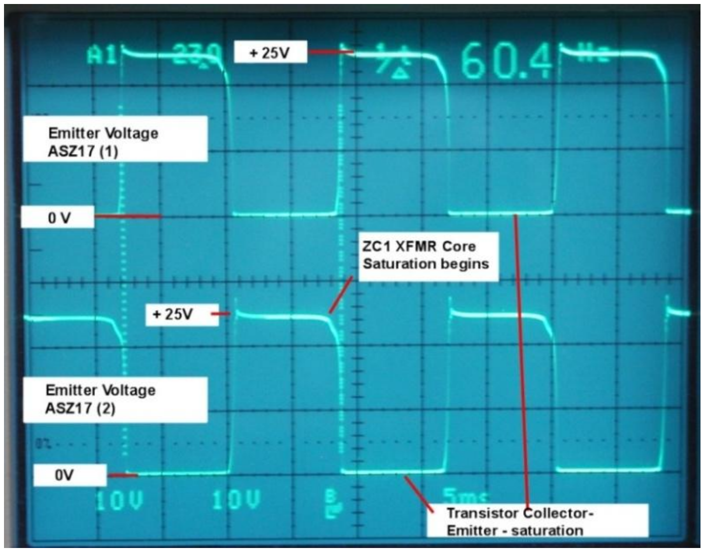
(The resistor power dissipation is 0.426 watts in each resistor so a 0.5W resistor is adequate however it is best to use a 1W resistor for reliability).

The top cover can then be fitted:

The photo below shows the unit working in a ZC1. It looks the part and suits the rugged character of the “ZC1 Machine”

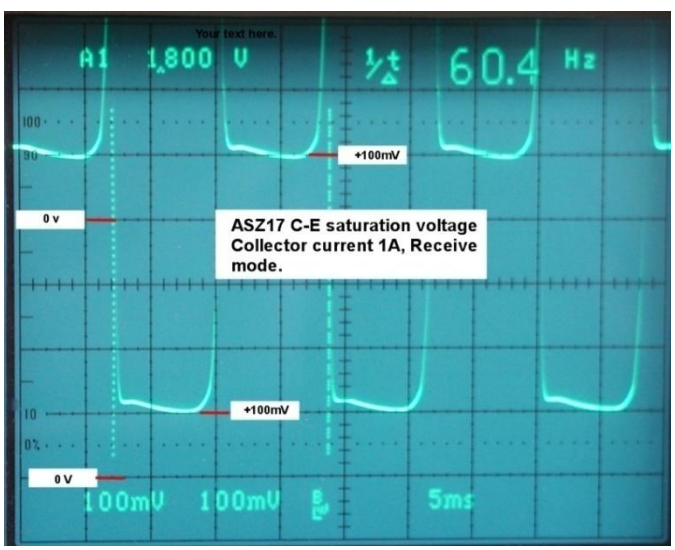


The following is a dual trace recording (using a Tek 2465b scope) of the emitter waveforms of the two ASZ17’s (these are the ZC1’s primary transformer connections) with the unit running in receive mode. As can be seen it self oscillates at close to 60Hz with a very clean switching waveform:



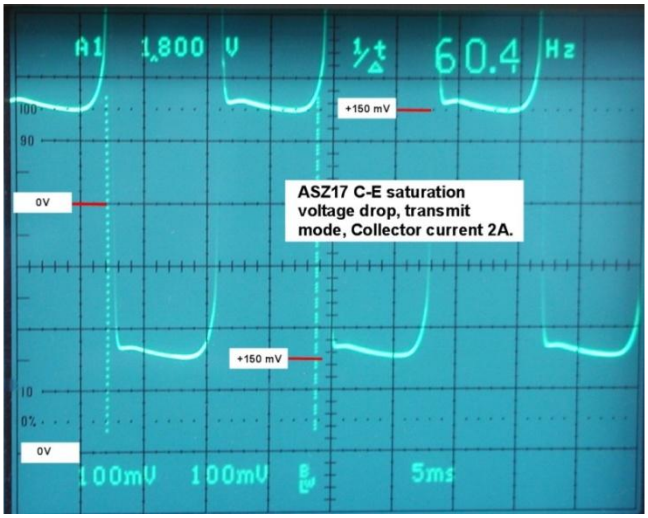
The labels on the scope trace were added to show the important features. Notice how the induced voltage which is 12.4 volts across one transformer primary side plus the 12.4 supply voltage (on this test) results in about 24.8 volts appearing on one transistor's emitter while the other transistor is conducting. Then after a time, due to magnetic saturation of the ZC1's main transformer core, the induced voltage fairly suddenly starts to fall. This initiates the conducting transistor to come out of conduction and the other to go into conduction for the next half cycle.

The base drive current for each ASZ17 transistor is in the order of 210mA and the collector current in receive mode is about 1 amp. To see how well the transistors were saturating, the scope gain was wound up on DC to 100mV/cm:



The recording above shows the wonderfully low C-E saturation voltage drop for the Germanium power transistor (ASZ17). It would be possible to achieve a low voltage like this with a low RDS on Mosfet, however as noted the ASZ17 Germanium BJT is very suited to this simple circuit.

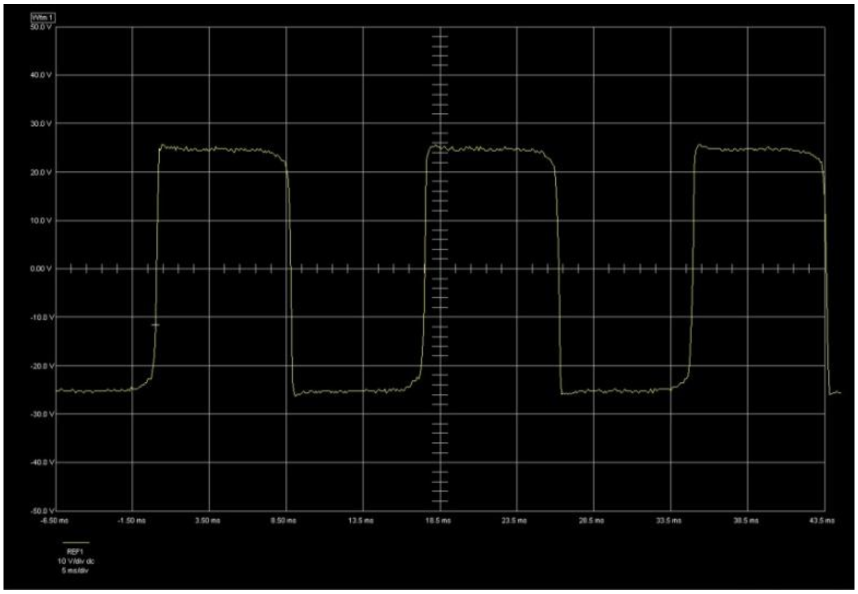
In transmit mode the collector current is about doubled to 2A and the saturation voltage increases a little to 150mA. (if these were a switching Mosfet this would correspond to an RDS on of 75 milli-Ohms by comparison)



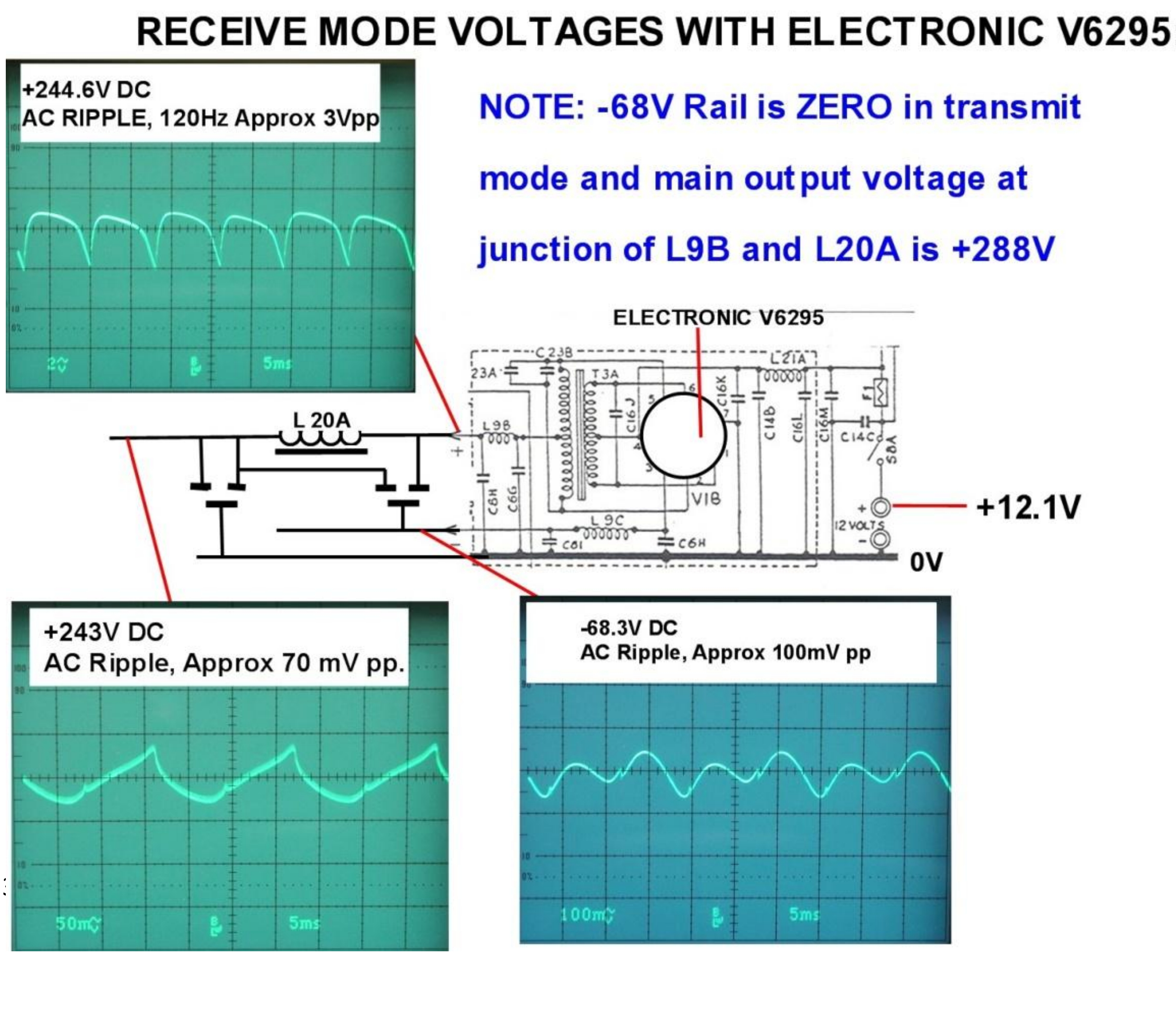
Notice on the photo above that the frequency cursors were left where they were so the scope still displays 60.4Hz, however the self running frequency has slowed just a little on the waveform itself due to the additional loading.

In transmit mode the power loss in the two ASZ17's transistor's collector-emitter circuit is about 2A x 0.15V = 0.3 watts. The base-emitter power is 0.21A x 0.45V = 0.09w, so the dissipation in each transistor is only about 0.4 to perhaps 0.6 watts as there are some additional losses during the switching transitions. So the whole assembly runs very cool on account of the size of the metal housing the transistors are screwed to.

In addition if an isolated scope is placed across the coupling transformer primary, pin 1 and 6, the following voltage waveform is attained and as can be seen is a 48v pp rectangular wave with a DC supply voltage of around 12V as shown below:



The diagram below shows the DC output voltages and AC ripple voltages in receive mode with 12.1V on the main input and the sender switch ON:



The output after L20A is very good, +243V DC with only 70 mV ripple. (As previously noted my ZC1 has 25uF filter capacitors, so with 10uF ones the ripple would be a little higher, but this is a very low figure for a power supply and tube equipment of this type).

Also as noted in the transmit mode the “ - 60V rail” is taken to ground, this elevates the output voltage on L9B.

Original ZC1 voltages from the Technical Description Manual:

Unfortunately the manual is non specific about the voltages in receive vs transmit mode and if in receive mode whether or not the sender switch was “ON” (this affects the HT voltage by about 12v). However some things can be deduced from the tables on pages 44 to 48. Firstly the highest LT voltage listed is 11.6V, so this suggests that this was the voltage at the fuse output connection. In receive mode the 6U7 IF (V1B) amplifier would be expected to have close to the main HT voltage on its anode, listed as 217.5V in the manual however using a restored original V6295 and 11.6V at the fuse output I get 225V.

Again with 11.6v at the fuse output and measuring this with the electronic V6295 yielded 235v, which is also the voltage measured at L19A.

This suggests that using the electronic V6295 results in a higher output than the original mechanical unit by about 235/225 or about 4% higher in receive mode (sender switch on). This is to be expected as the mechanical unit can’t quite reach as close to a full 50:50 duty cycle due to its contact gaps and the time that neither contact is closed.

Although not specifically stated, V4B (6V6 modulator tube) Screen Grid voltage is listed at 235V in the table and its anode at 227.5V. Presumably, these voltages are measured in Transmit mode, while transmitting and are higher as the -60V rail is grounded. Measuring these (again with 11.6V at the fuse output) yields 267V and 262V respectively. Therefore, in transmit mode the output voltage from the electronic unit is roughly about 14% to 15% higher than the original unit under test conditions in the ZC1.Other tests into dummy loads totalling 0.07mA gave about a 6% improvement. In any instance this electronic unit is superior to the electro-mechanical V6295.

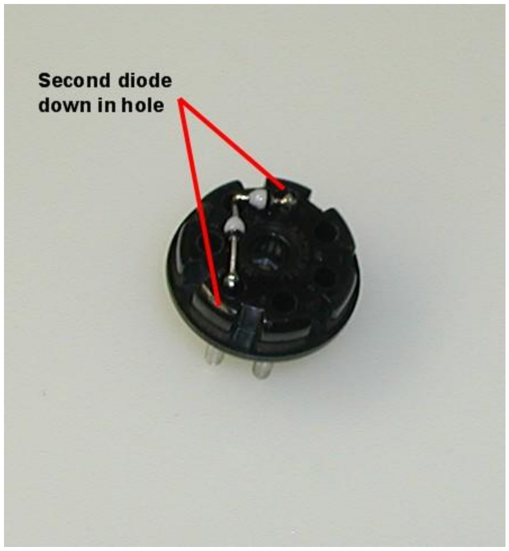
3) THE MOSFET V6295 OPTION (Oscillator Driven):

It can be a difficult task for most home constructors to wind transformers, although they can be ordered from transformer winding companies. Also the mechanical housing of the ASZ17/ Transformer electronic V6295 needs the services of a machine shop. Therefore to help ZC1 enthusiasts I have designed three other alternative versions. The one below is based on Mosfets driven by an independent oscillator with hardware & components that are readily available. The brass plate and brass wire (made by K & S engineering USA) is available from Mr Toys in Australia. The UX7 base is a standard American Amphenol part, available on Ebay usually.

Other parts are available from Jaycar Electronics or RS Components or Element 14.

Any V6295 replacement still requires some type of chassis or skeleton to support it and preferably a metal mounting for the output devices. The simplest way to do this is to start with a standard Amphenol UX7 plug and fit it with a structure composed of a brass spacer, brass plate and a grounding wire passing into pin 7 of the UX7 socket. The basic parts are shown below. To assist making sure the 3mm diameter hole in the plug is drilled on centre a temporary 3mm spacer can be placed in the ¼" recess to guide the drill. The hole is then countersunk from the pin side of the plug.

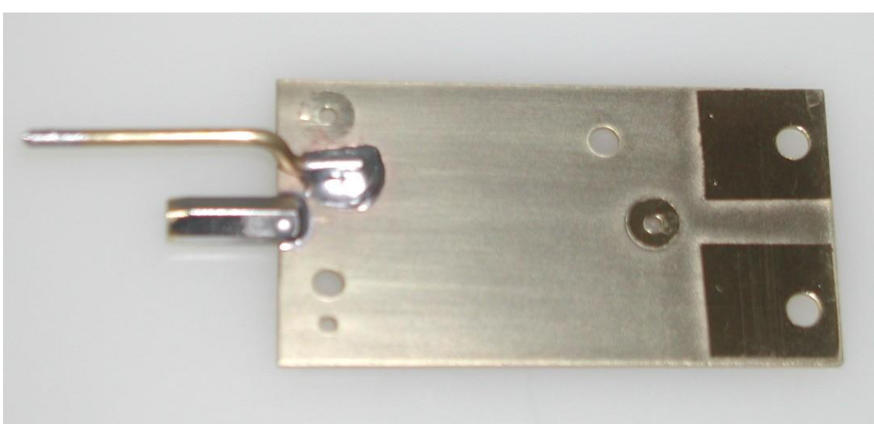
The four BY448 rectifiers are then placed on the plug assembly:



The 3mm thread x 15 mm nickel plated hex brass spacer has a 2 to 2.5mm deep slot cut in it to accommodate the brass plate. To do this I used a junior saw and a fine flat file. The spacer is made a push fit the easily soldered to the plate by holding it with grips over the flame of a gas stove or with a suitable soldering iron.

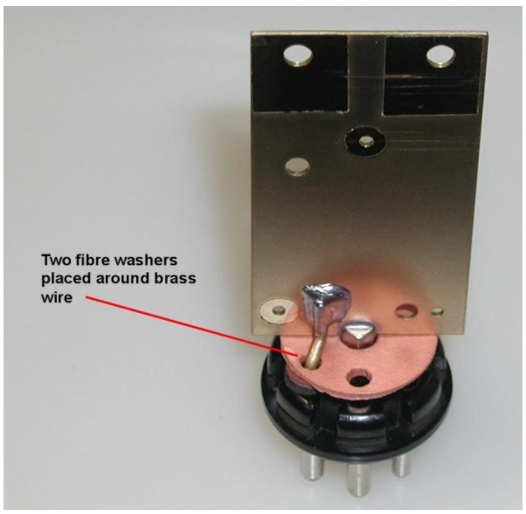
The brass plate can have its holes drilled before or after fitting to the spacer, but it might be a little easier to do it first because the plate sits flat.

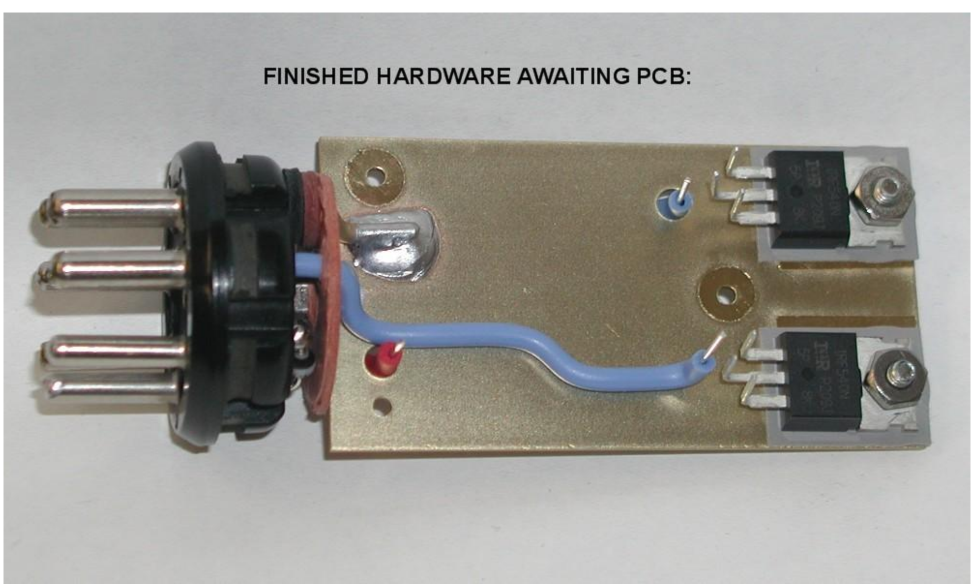
The brass plate & spacer can then be temporarily fitted to the plug to align it parallel with the ZC1's vibrator transformer's face when the unit is plugged in and a brass wire positioned to pass from pin 7 (earth) of the plug to the brass plate, and the brass wire then soldered to the plate. The photo below shows the finished metalwork.



The spacer required rounding off a little at its end to fit into the deep hole in the UX7 plug. The assembly was masked with tape where the power Mosfets and PCB spacers go to allow a good earth connection & then it was sprayed with lacquer accounting for the appearance in the photo above.

The hardware is then further assembled to be ready to receive the PCB. A cardboard washer, one inch diameter, made from Jaycar electrical cardboard is used to cover the rectifier connections and the other wires are 0.7mm diameter tinned copper, covered in silicone rubber insulation, but PVC would be ok:

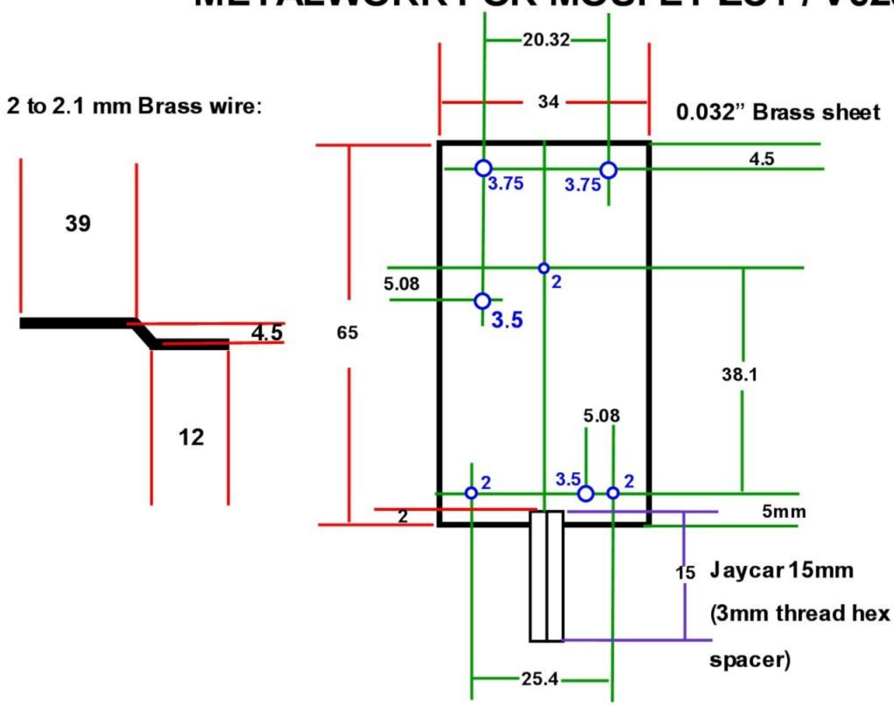




The thick (2mm diameter) brass wire ensures that the plate cannot rotate easily even if its fixing screw becomes loose and it gains a solid earth connection.

The hardware dimensions are shown below. The hole spacings are all multiples of 2.54 mm because the PCB is laid out on a 2.54 mm grid. The hole diameters are in blue:

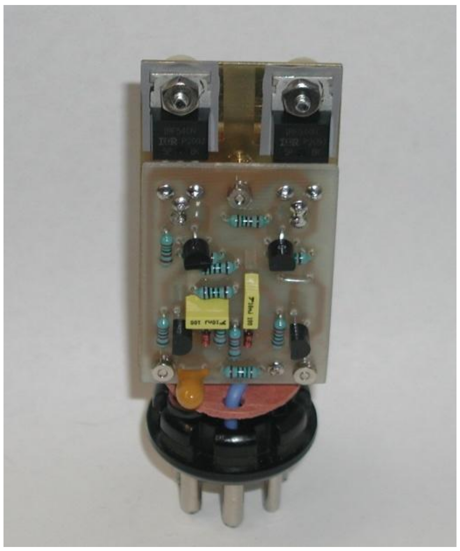
METALWORK FOR MOSFET ZC1 / V6295.

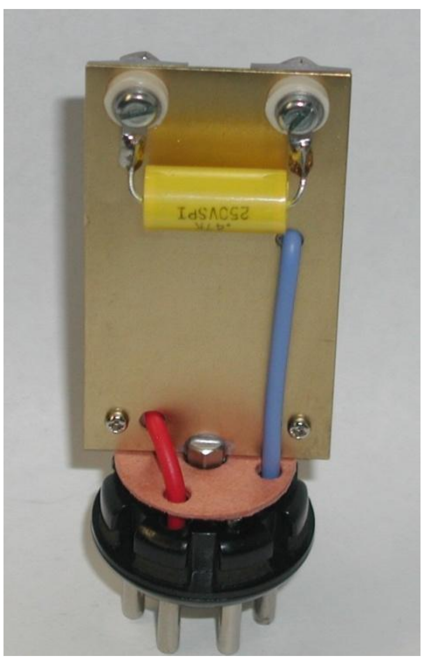


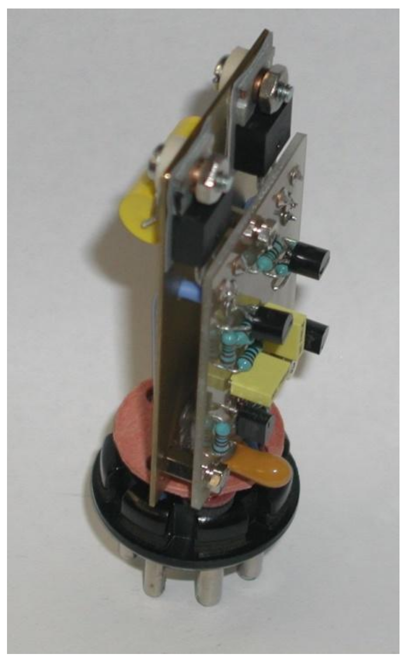
FRONT (PCB MOUNTIMNG SIDE VIEW) Dimensions in mm.

The PCB is mounted on three 5 mm long 2mm internal diameter nickel plated brass spacers.

The PCB is shown below. This could also be hand wired on plated through hole spot board if making a PCB was difficult. The PCB shown here was made with Jaycar iron on film applied with a clothing iron at 140 degrees C, for four minutes, then etched with Ferric Chloride. 1.6mm diameter brass eyelets were added where the transistor pins and wire connections pass through, however a professionally made PCB would/should have plated through holes here instead.



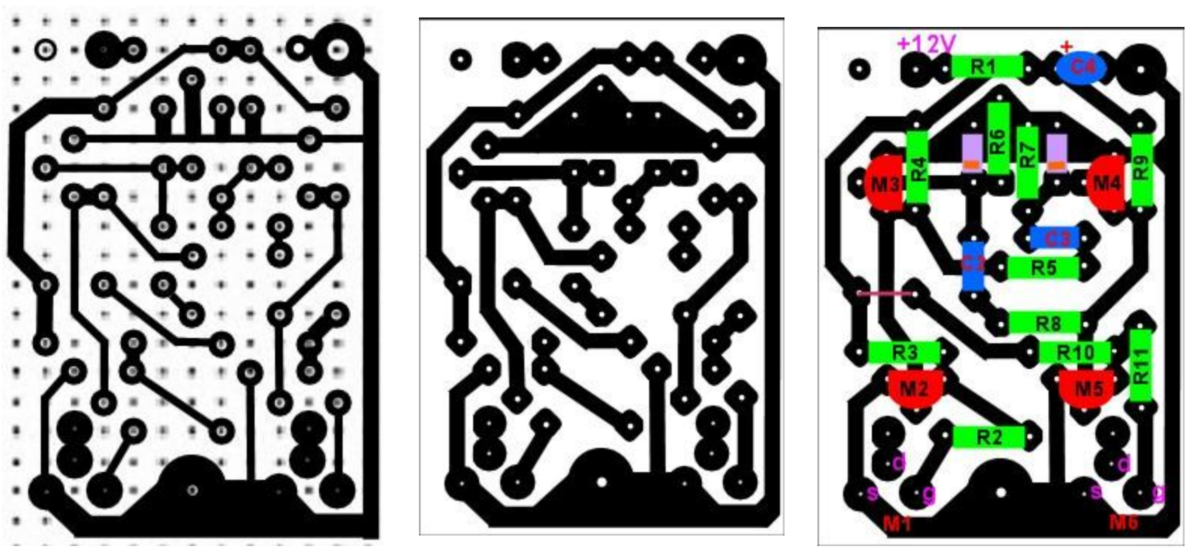




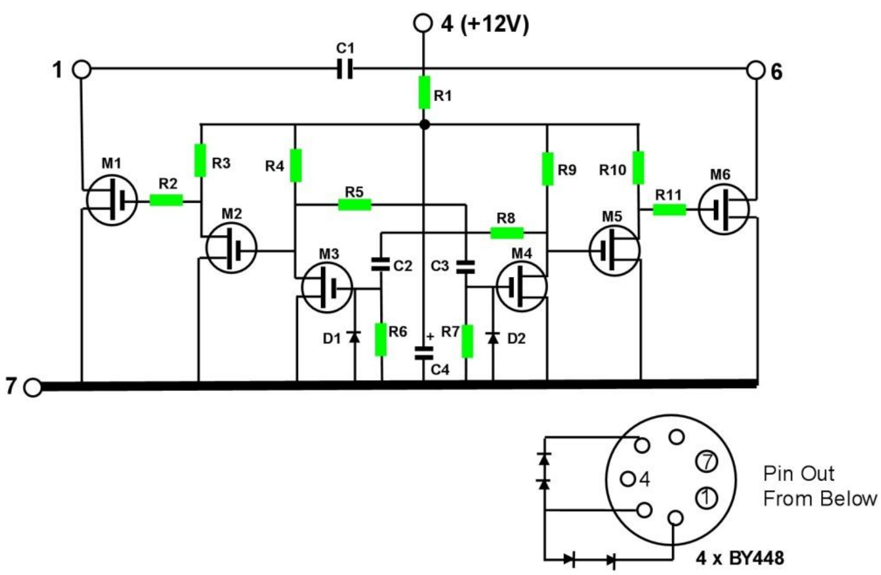
A “tuning” capacitor (C1) is added between the Drain connections of the IRF540N Power Output Mosfets. (The purpose of this is explained in detail later).

PCB layout and Schematic:

The photos below shows the PCB layout tracks viewed from PCB top. All the holes are placed on the 2.54 mm grid locations, except the negative leg of C4. The centre photo is the actual image used for making the pcb. The tracks and pads, unlike most modern pcb’s are kept as large & wide as possible as fine tracks can be difficult to achieve at home. The PCB is approximately 33.5 x 47mm size.



ZC1 MK2 MOSFET V6295 REPLACEMENT. H. Holden, 2013.



Note: In the schematic above the mosfets are drawn with a generic symbol and without their internal Drain-Source diodes. All the mosfet's source electrodes connect to ground (negative), or pin 7.

Parts List:

(all resistors ¼ w but could be 1/8 watt also)

R1 = 150R

R2, R3, R10, R11 = 1k5

R4, R9 = 10k

R5, R8 = 100R

R6, R7 = 270k

M2, M3, M4, M5 = BS270

M1, M6 = IRF540N

D1, D2, = 1N4148

C2, C3 = 10nF, 100V MKT.

C4 = 15 uF 35V Tantalum.

C1 = 0.47 uF, 250V poly.

Circuit operation:

As can be seen the circuit is based on a multivibrator with two BS270 mosfets, M3 & M4. This zero bias configuration gives more reliable starting from low voltages than biasing these mosfets to an ON condition analogous to the usual transistor multivibrator.

Due to the high impedance at the gates, a large value gate resistor and small value timing capacitor can be used (270k & 10nF) which gives accurate timing and avoids the use of poor tolerance electrolytic capacitors in a typical transistor low frequency multivibrator.

Diodes D1 and D2 clamp the gate drive signals to -0.7v below zero. The multivibrator runs close to 110Hz, similar to the original V6295.

Due to the fact if anything stops the multivibrator, or for some reason it didn't start due to a very slow rising supply voltage then the drain potential of M3 and M4 would go high, which would be problematic if these were used to drive the output mosfets directly. Therefore an inverting buffer stage is included M2 and M5 which also helps to isolate the multivibrator from the output stage. The supply to the multivibrator is also heavily filtered with the 150R resistor & 15uF capacitor to ensure the voltage transients which are significant on pin 4 (+12V supply) do not cause premature triggering of the multivibrator when it is in a vulnerable condition about to change state.

Separate rugged mosfets (the BS270) were used rather than a Cmos IC here, as the former have much higher voltage ratings (60V) and are much more immune to damage from spikes & transients and do not require so much protection on the power supply feed as a Cmos IC does. This circuit will start from voltages as low as 6V.

It is normal practice in switch-mode power supply design to drive the gates of the output mosfets from a low impedance source, typically 100 ohms or as low as 10 ohms. The power mosfet's gates often have a significant capacitance in the order or 500 to 5000pF depending on the mosfet type. If the gate series resistor is too high in value it can slow the switching time down and decrease the efficiency (increase the mosfet's heating) because it spends longer time in an "intermediate state of conduction" rather than on or off. Also the switching frequency in often in the range of 20KHz to 100KHz in switch-mode PSU's, so there are many switching events per unit time and these losses add up.

In addition, switch-mode power supply transformers are generally wound with a low leakage inductance, often with ***bifilar wound*** primary windings….. however, the ZC1 power transformer is not like this at all, it has a relatively high leakage inductance between the halves of the primary windings and also operates at a much lower number of switching events per unit time than a typical modern switch-mode power supply.

Therefore, the design rules for this application are a little different to a modern switch-mode supply. In this instance extremely rapid switching of the output mosfet is disadvantageous because the transformer's primary winding leakage inductance (and leakage reactance) is so high that this produces very high voltage transients on the contralateral (other) or fellow mosfet's drain at the moment one mosfet switches on, in the order of 70 to 100v and at a resonant frequency at about 50KHz.

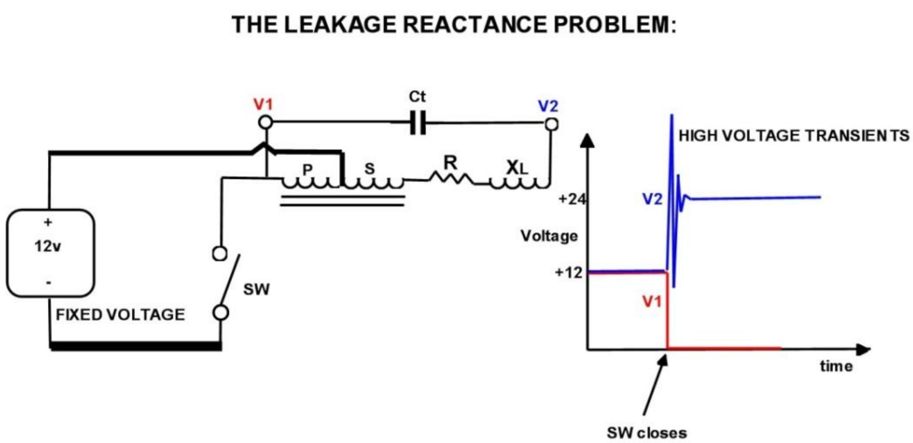
This is ameliorated a little by the 1k5 gate drive resistor network which forms a mild LPF (low pass filter) with the IRF540N's gate capacitance. Also the added 0.47uF "tuning" capacitor lowers the resonant frequency of the leakage inductance – capacitance network to about 20KHz and reduces these voltage transients on the mosfet's drains to a low acceptable level when switching occurs (more about this below).

THE LEAKAGE REACTANCE PROBLEM:

It is worth looking at the leakage reactance problem and how the voltage overshoots come about. If these overshoots (oscillations) are too large they can exceed the Drain-Source voltage of the mosfet (or the Collector-Emitter voltage if a transistor is being used) depending on their ratings. These spikes can be a potential source for insulation breakdown of the transformer windings or tuning capacitor/s failure.

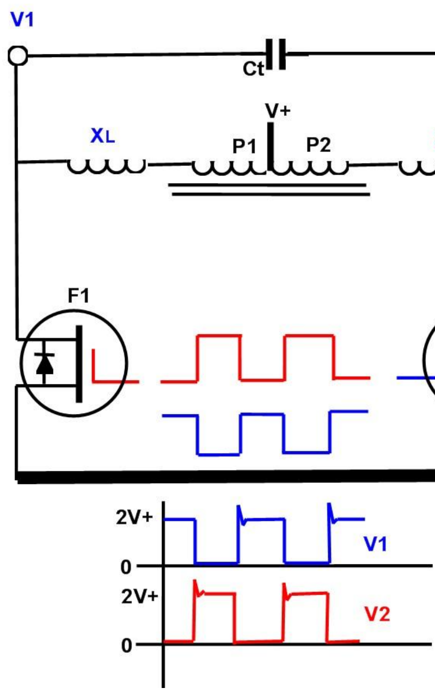
The diagram below illustrates a “single ended example” where a transformer with two windings, a primary P and a secondary S is initially at rest with no winding current, then a switch closes. The switch could represent a mosfet or transistor, and the two windings could represent the centre tapped primary winding of the ZC1’s vibrator transformer for example (even though they are both the “primary” in actual use, one acts as a “secondary” in terms of the transients generated).

X_L is the leakage reactance (due to an inductance L acting in series with the windings) which comes about due to the fact that not all of the magnetic field links both windings P & S. The leakage reactance appears in series with the primary, or the secondary winding when the other winding is shorted out *or has a fixed voltage applied to it*. The tuning capacitor Ct could be the winding self capacity plus any other added capacity. The resistance R is that of the ohmic losses of the windings mainly:



When the switch closes (initially at least before there is time for core saturation) 12v DC is applied to the primary winding P, effectively shorting it out from the AC perspective so the leakage reactance X_L appears in series with the secondary winding S. The induced voltage is of a polarity that attempts to raise V2 to 24V as one side of secondary winding S is connected to +12V. However to achieve this Ct must be charged, and it forms a resonant circuit with the leakage reactance X_L, with some damping by R. Therefore oscillations (spikes or ringing) occur on terminal V2. The frequency of this resonance is largely determined by the leakage inductance X_L, and the tuning capacitance Ct, and the resistance R plays a part as the damping is fairly heavy, but there can often be 4 or 5 cycles of oscillation or “Ringing” before they damp out. This is why increasing the tuning capacitance, lowers both the frequency and the amplitude of these oscillations or ringing. To look at it another way; the Q of this resonant circuit comprising R, X_L and Ct is lowered with a larger tuning capacitor because the resonant frequency is down shifted and the inductive reactance of X_L is lower at that lower resonant frequency.

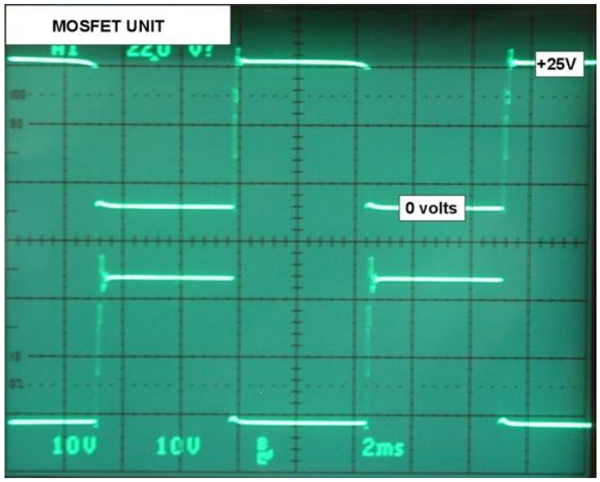
In the case of the push-pull rather than single ended example above, the same situation occurs as shown in the diagram below, the resistance is omitted for clarity:



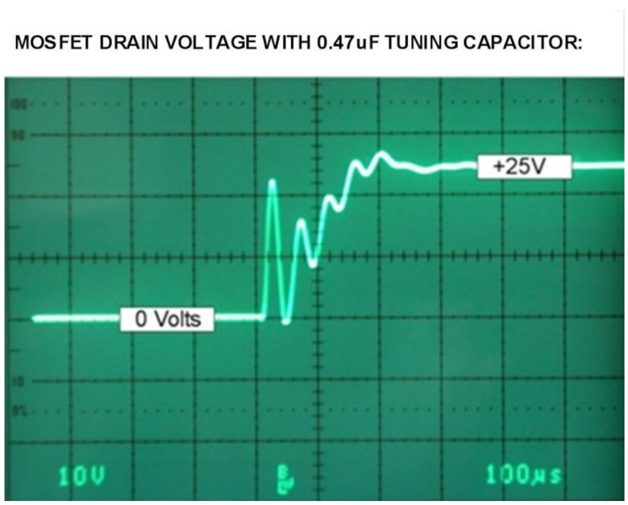
When FET F1 switches on (red drive waveform high) voltage **V1** is taken rapidly to zero in a few microseconds or less. The value **XL** vanishes when F1 is conducting as a fixed voltage +V is applied to winding P1 and all the leakage reactance then appears as **XL** in F2’s Drain circuit. F2 is also off at this time. Ct is in resonance with **XL** so the leading edge of the voltage **V2** has the ringing & overshoot. The situation is reversed when F2 conducts, making **XL** vanish and placing all the leakage reactance **XL** into F1’s drain circuit.

Also noted here that the amplitude of the mosfet drain waveforms or power transformer primary connections are transformed to twice the supply voltage, which holds true until the magnetic core of the power transformer starts to saturate. For the ZC1 transformer this is about 8mS, so using a 100 Hz drive waveform does not take it near core saturation, unlike the ASZ17/Transformer method which relies on core saturation to sustain oscillations.

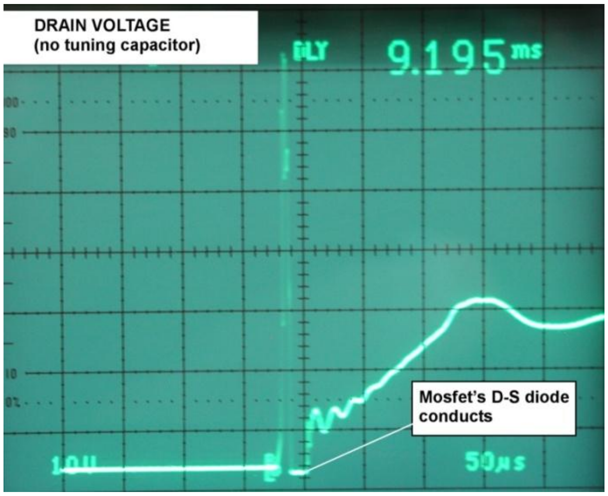
The following is a recording of the ZC1's primary winding voltages with the completed mosfet unit running in the ZC1. The oscillations are visible on the Drain connections (transformer primary) immediately after one mosfet comes out of conduction and the fellow mosfet goes into conduction. The mosfets switch very fast, less than a few microseconds even with the 1k5 gate resistors.



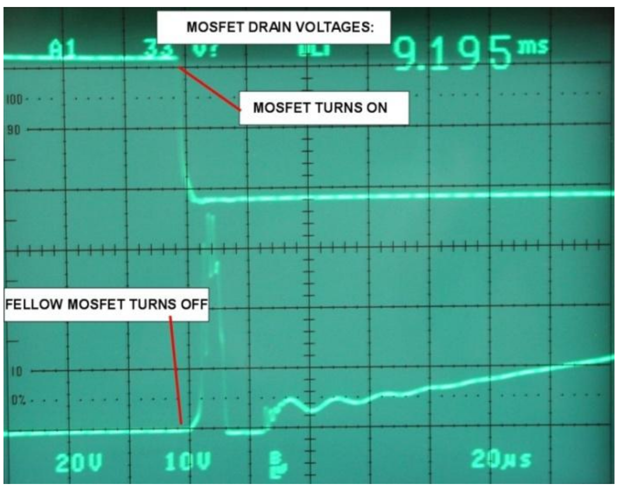
To get a closer look the photo below shows the area of ringing with the scope's delay time-base, with the 0.47uF tuning capacitor the ringing frequency is about 20KHz:



Without the added tuning capacitor (photo below) the ringing frequency is about 50kHz and the peaks are much higher, and are a threat (other smaller oscillations are seen superimposed due to the ZC1's vibrator transformer high voltage secondary windings and their leakage reactance and associated capacitance too).



As can be seen in the recording above, the initial peak is very high around 70v and on its negative half cycle actually causes the mosfet's internal Drain – Source diode to conduct, clamping off the negative half cycle. Another photo shows the timing of this transient which occurs just after the mosfet switches on and its fellow turns off.

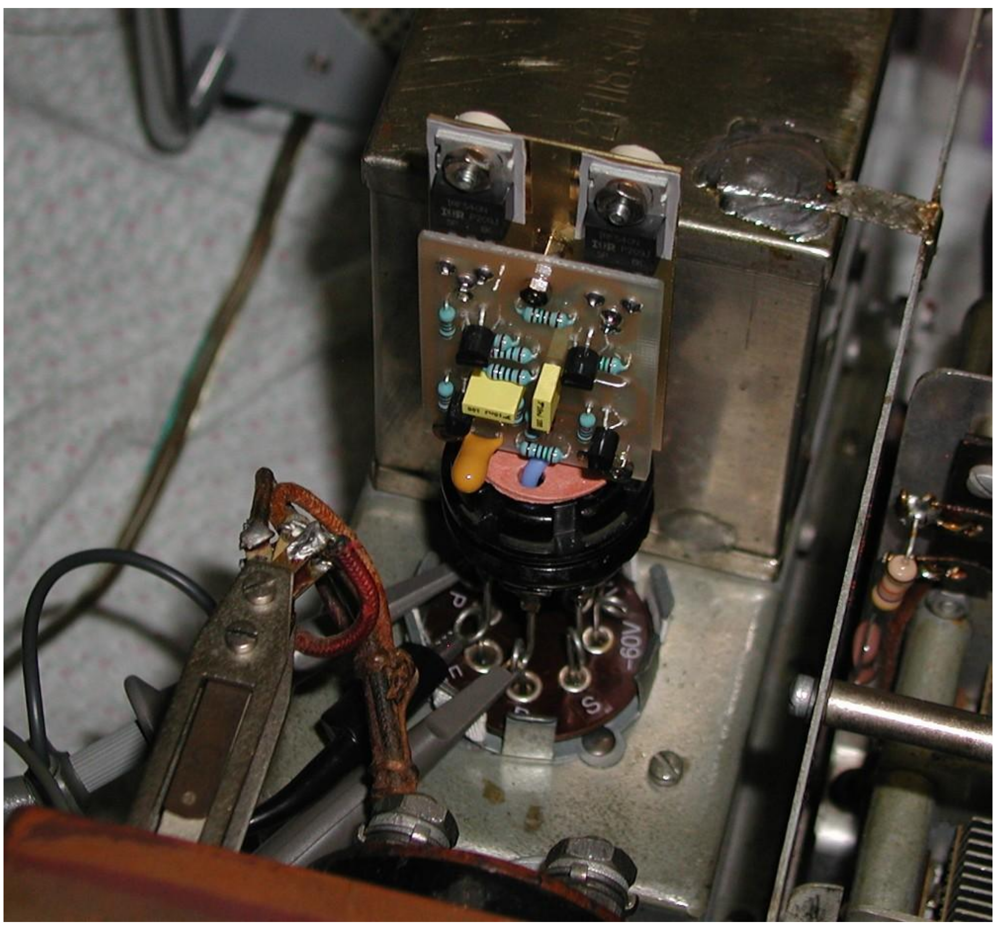


Therefore it is important with this mosfet version to incorporate the added 0.47uF tuning capacitor, and for any version using silicon transistors driven by an ***independent oscillator*** (like commercial transistorised units) it would be wise to include the capacitor when used with a ZC1's vibrator transformer.

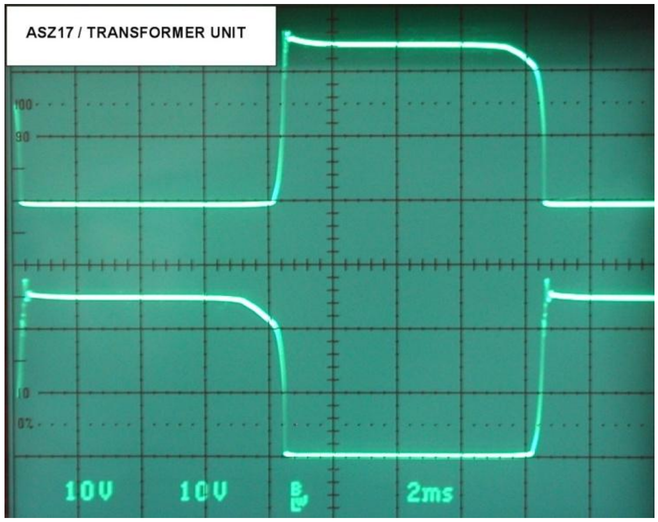
(With the mechanical vibrator this first peak is lower at around 30 to 40V. This is because with the reduced duty cycle the transformer's primary voltage falls from 24V to about 16V prior to the next switchover as the duty cycle is lower with the oscillating contacts and the energy transfer to the circuit comprising Ct and X_L is a little lower).

Another potential method to solve the leakage reactance/voltage spike issue is to snub off the high voltage transients with a TVS (transient voltage suppressor) to around 30V however there is a little more chance of RFI with this method versus tuning the resonant frequency downwards with the added tuning capacitor. A bidirectional 30 to 40V TVS placed between the output mosfet's drains would work.

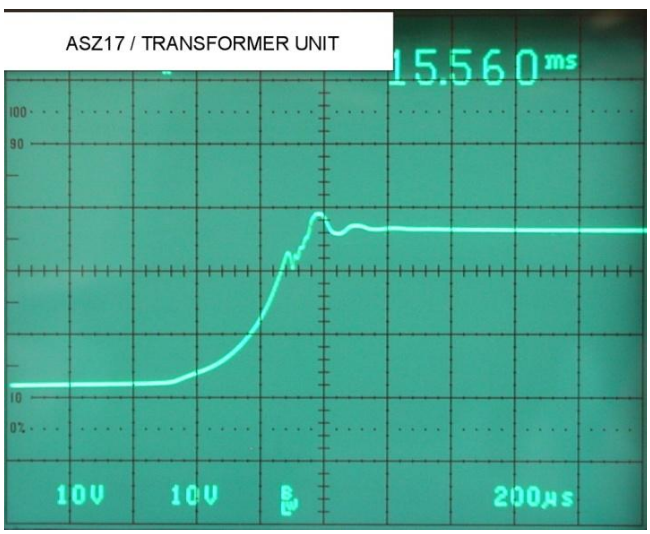
The following photo shows the completed mosfet unit, under test, running on the test extension tool in the ZC1. It is a good performer and there is no significant RFI unlike the original mechanical vibrator:



NOTE: In the case of the transformer driven unit with the ASZ17's, the additional tuning capacitor is **not** required. The picture below shows the switching transients with this unit and the standard ZC1 vibrator pack:



And the delay time-base view:



The likely reason why the ASZ17/Transformer circuit does not have such severe ringing and overshoot is that prior to the switching event the transformer core is entering saturation and this lowers the leakage reactance and gives a smooth switching transition with minimal overshoot as can be seen from the above waveform.

DARLINGTON V6295 OPTION:

This electronic V6295 version seeks the simplicity of the self oscillating ASZ17/transformer version, but without the transformer and a more easily crafted housing made from brass rather than from aluminium.

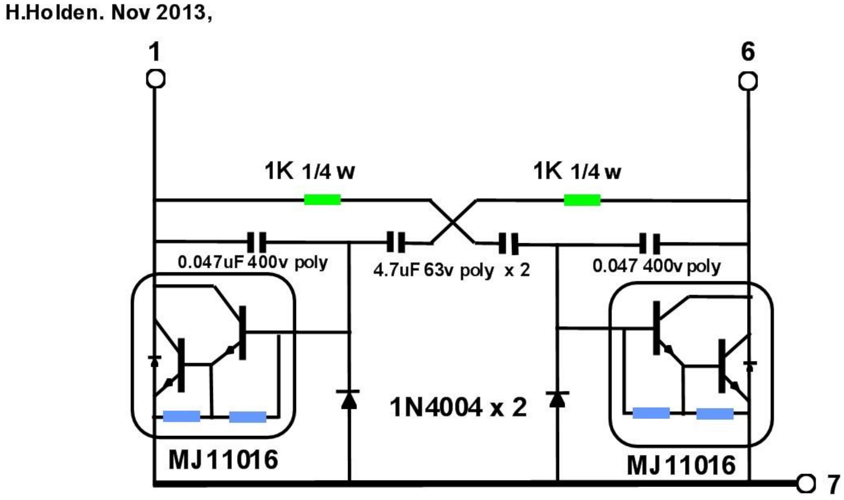
If a self oscillating **transformer-less** version of the V6295 were to be made then the first step would be to use output devices with a lower drive current since the drive power to one transistor's base-emitter circuit would have to be taken directly from the (fellow) or contralateral transistor's collector circuit which has a potential of 24V to 25V during switching. A single BJT (like the ASZ17 or silicon 2N3055) would be unsuitable as the base current required is 0.2A and if this is acquired from the fellow transistors collector circuit (peak voltage 24V) then the total base drive resistor's power dissipation (for both transistor together) is excessive at nearly 5 watts. This leaves options of IGBT's, Mosfets and Darlingtons as possible suitable output devices. The Darlington transistor has the advantage of a low input threshold voltage (1.4V) so the circuit can be made to self start (oscillate) from low power supply voltages, in fact the Darlington circuit described here starts oscillating at a supply voltage as low as 3 volts. The Darlington power transistor also has the advantages of internal base resistors and collector emitter diodes saving on parts.

Frequency response limiting and stable self switching can be obtained with either electrolytic capacitors of 16uf to 20uF with series 1.5R resistors to form a snubber network on each transistor's collector to emitter (ground), or alternatively smaller value 0.047uF Miller integrator capacitors from the collector to base circuits of the Darlington transistors. Without this negative feedback and using these very high frequency capable Darlington transistors the oscillator circuit is highly unstable and oscillates at a high frequency corresponding to the main ZC1 power transformer's primary leakage reactance and associated capacitances in resonance. When this happens the transistors overheat and are destroyed if the condition is allowed to persist.

Also by using Darlington transistors as switching elements the base drive power is reduced by a factor of 10 to 20 compared to a single BJT. The positive feedback capacitors to sustain oscillation from collector of one transistor to the base of the fellow transistor can be a modest value of 4.7uF and be of a non electrolytic type. Electrolytic capacitors are better avoided where the exact values are responsible for setting a time constants(The drive power can be further reduced using a self oscillating Mosfet circuit).

The following circuit shows the transformer-less self oscillating V6295, based on MJ3001 or MJ11016 NPN Darlington transistors. The oscillation frequency (62Hz) is nearly identical to the transformer coupled version in Receive mode on the ZC1.

SIMPLE V6295 TRANSFORMERLESS OPTION:



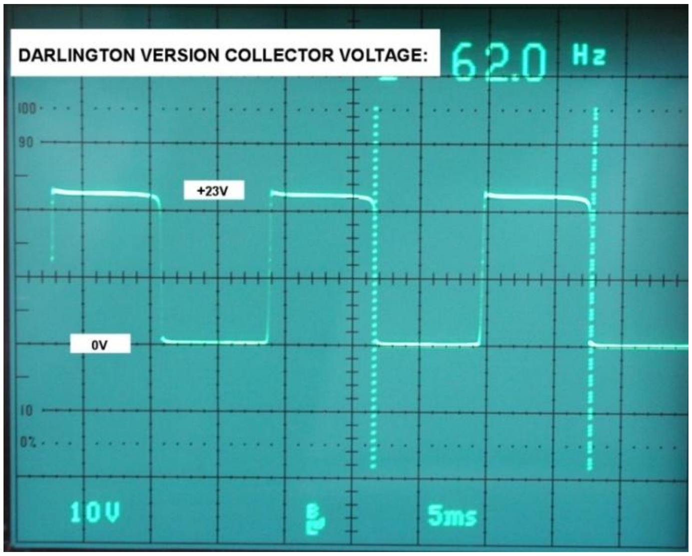
Again the numbers are the pin numbers of the UX7 socket (The arrangement of the 4 BY448 rectifiers is the same for all electronic V6295 versions).

The waveforms below show that the ZC1's transformer switching voltage is practically identical to that for the ASZ17/transformer version. One difference is that the collector-emitter saturation voltage drop of the Darlingtons in this application with a peak collector current of 2A is about 0.9V, versus the 0.15V for the ASZ17's. Therefore this Darlington unit results in an output ZC1 power pack of about 6% less than the ASZ17/Transformer unit. However, the output voltage and efficiency is very similar to the original electro-mechanical V6295. The advantage is that the Darlington unit is "transformer free" and is simple for the home constructor to manufacture.

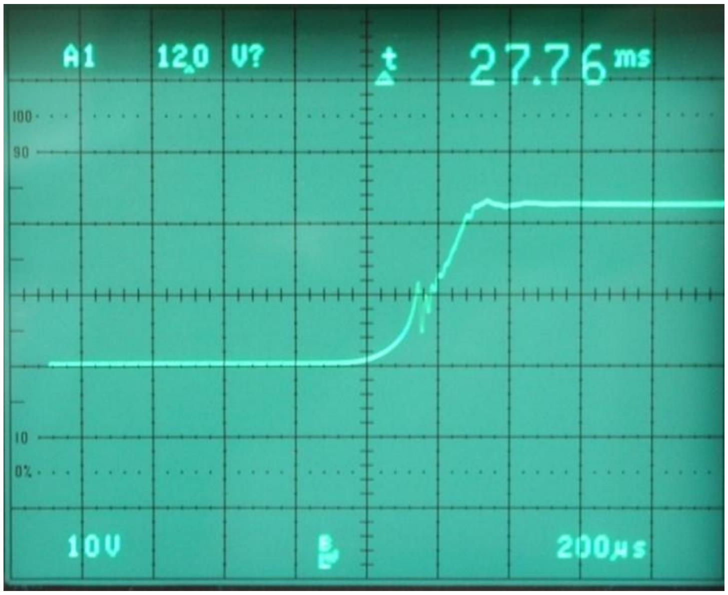
RS components have the MJ11016G as a pair in a box, part 463-000 for around $5 for the pair which is very good value for two such TO-3 transistors. The manufacturer is ON Semiconductor. Another suitable part is the MJ3001 which is available on Ebay and also works fine in this circuit.

Darlingtons are a reasonable choice for a **12V system**, even though in this application the output voltage is 6% lower than for the single BJT transformer version – (see analysis page 62 about when Darlingtons are suitable or not).

The photo below shows the collector waveform for the Darlington unit:

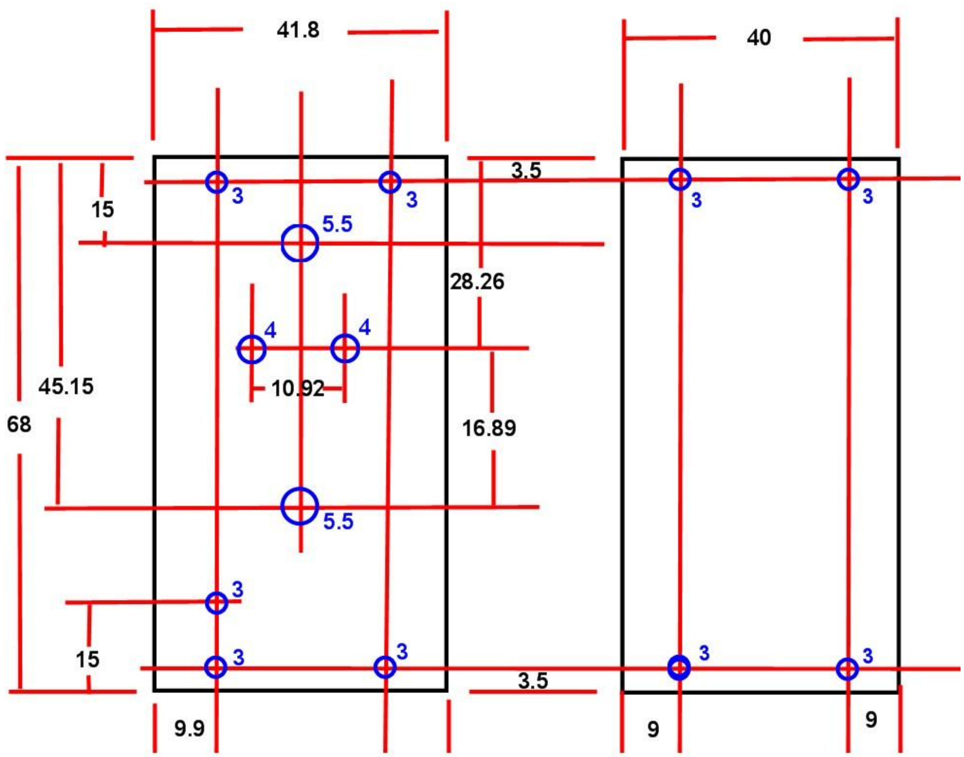


And the delay time-base view below to look at the details of the voltage rise and overshoot, showing only a small resonance during the switching event with no significant collector voltage overshoot. Due to the 0.047uF Miller feedback capacitors and the operating frequency being close to 60Hz. (The 0.47uF tuning capacitor required in the 110Hz mosfet unit was **not** required here).

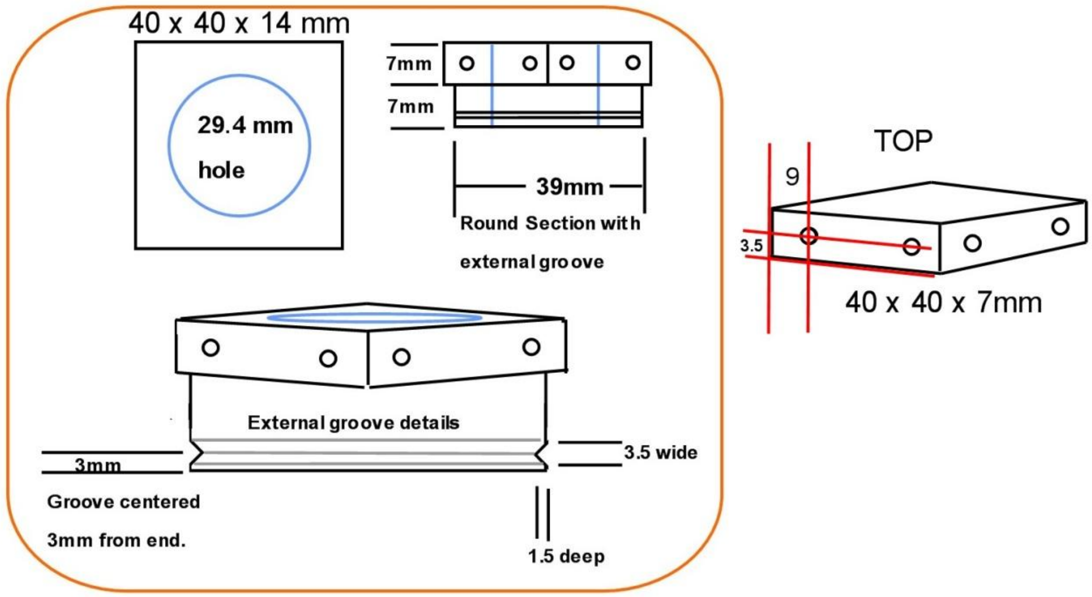


BUILDING THE DARLINGTON TRANSFORMERLESS VERSION.

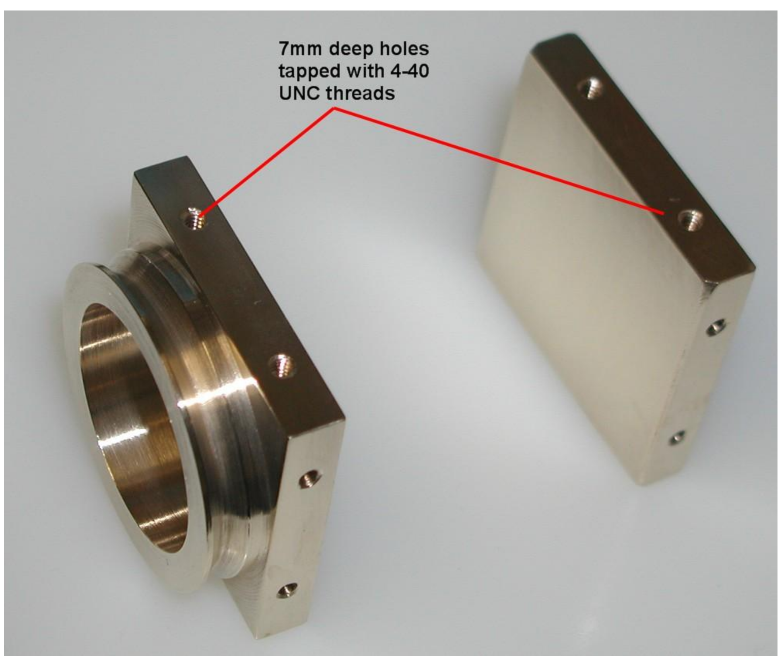
Four Brass plates (two of each type shown below) are prepared with the following dimensions. When working with 0.8mm thick brass plate it is best to mark & drill 1mm pilot holes first and drill the holes out one size step at a time to get to the final size. The 0.032 inch thick brass plate is made by K&S engineering and is stocked in Australia by companies selling Models such as Mr Toys.



0.032" (0.813mm) BRASS PLATE

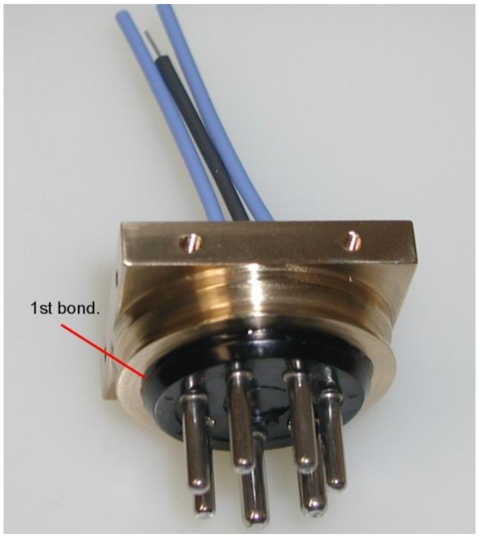


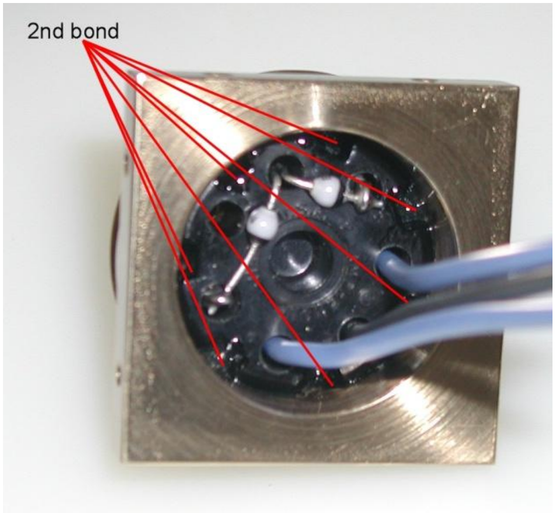
The machined brass base and top are shown above. These were turned up by a local machine shop. I added the threaded holes. The holes are 7mm deep. One point on taping into blind holes: Use a taper tap first and lubricate it with WD40 during the process. Then wash all the swarf out of the hole with a jet of contact cleaner from the applicator tube. Then tap to the base of the holes with a bottom tap to ensure the thread runs all the way to the hole's base. Then again wash out the swarf with the contact cleaner. Obviously it is important to be patient and careful when marking and centring and drilling the holes which are all 9mm from the edges of the square section as per the diagram above.



The Amphenol 7 pin plug base is prepared as usual with the BY448 rectifiers. Only three wires (the two collector wires and ground) are required as the +12v connection (pin 4) is not used.

This plug arrangement is simply glued into the brass base. This is best done as a two step procedure where a small amount of 24Hr Araldite is use to attach it and align in on the correct axis when the unit is plugged in. When that is dry, more araldite can be added to the well created by the edges of the plug and the inside of the brass housing without the risk of it draining out before it sets because the 1st bond has sealed it:





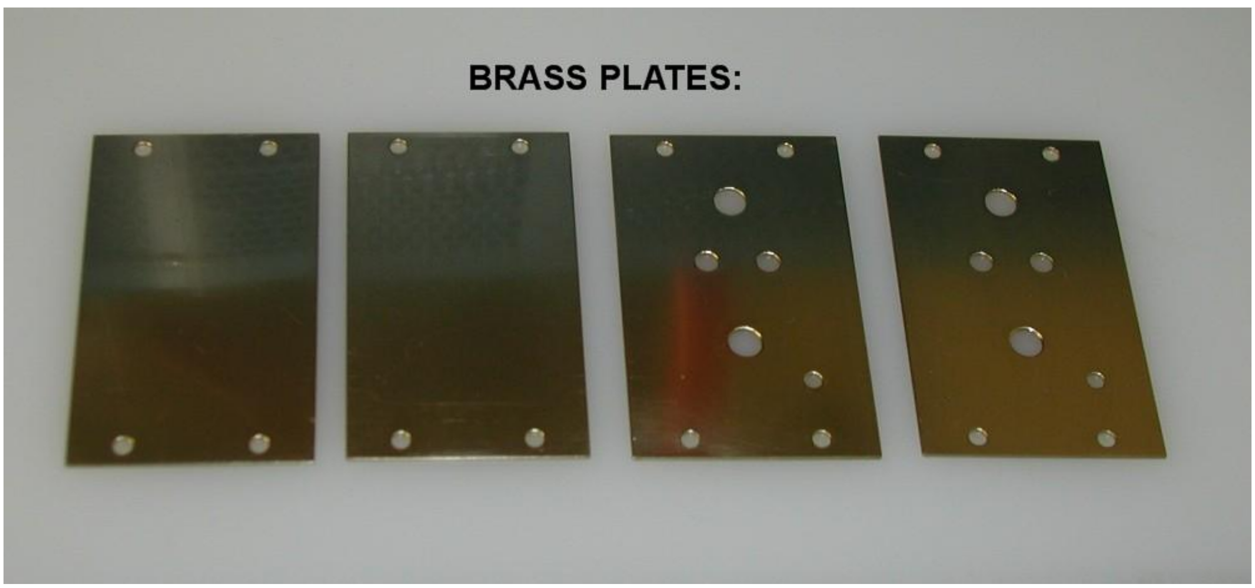
The photo below shows the units being assembled. In this case TO-3 transistor sockets were used. There are made by AUGAT in the USA. They are usually available on Ebay.

I would recommend them because they create convenient tie points for components, obviate the need for insulators, nuts/washers & lugs for the collector terminals and the transistors are easily removed for testing, replacements and experimental purposes and the transistors don't have to be soldered to either.

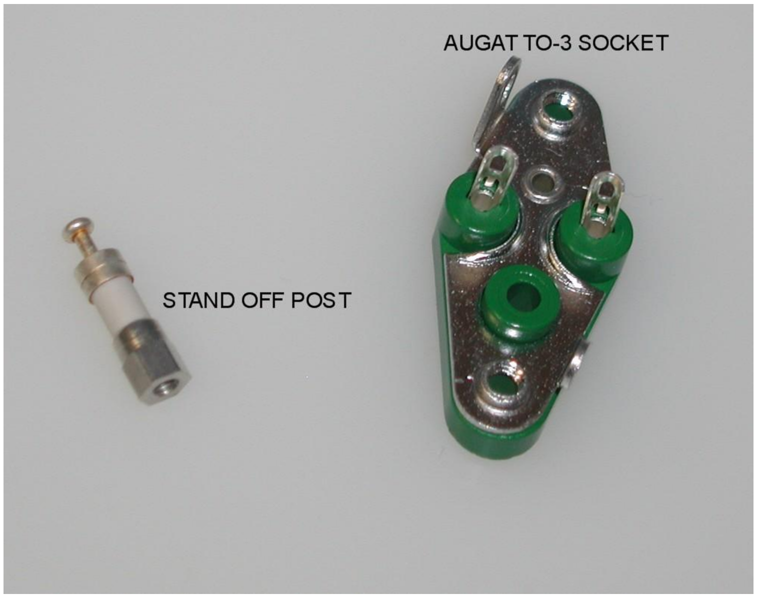
(Obviously the unit could be made without these sockets with direct wiring to the transistor legs and the usual TO-3 mounting gear. If the AUGAT TO-3 transistor sockets are not used and the transistor mounted with the usual insulator set, then the 5.5 mm holes in the brass plates should be 4mm).

(It is a sad fact that the TO-3 transistor housing is now obsolete and it is easy to forget that epoxy cased flatpack transistors have always had slightly inferior maximum temperature ratings to all metal style transistors like the TO-3).

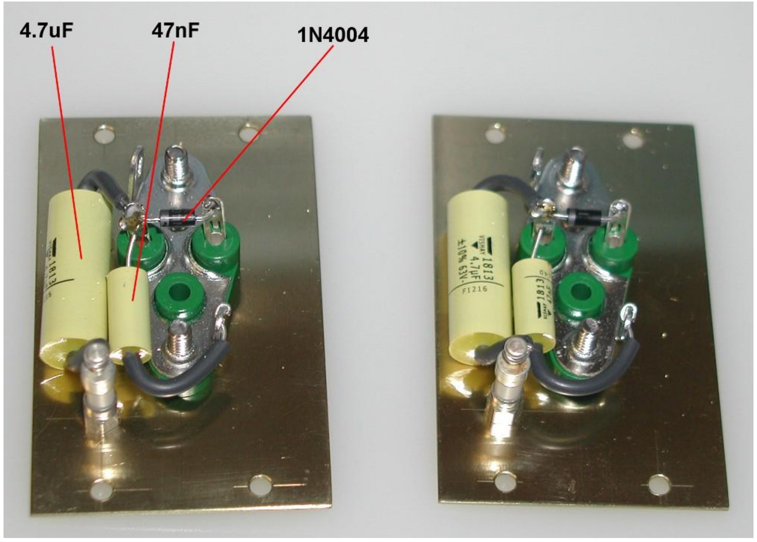
The four brass side plates for the unit are shown below:



The Augat transistor sockets and an insulated mounting post are shown below. These single type of insulated mounting posts are becoming rare. Surplus Sales of Nebraska still stock a range of mounting posts like this. Another option is a phenolic tag strip with a single lug:



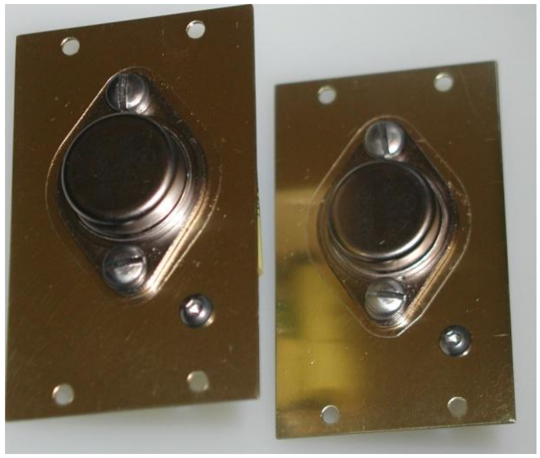
The photo below shows the capacitors & diodes mounted to the socket & post. Each left or right side panel is identical.



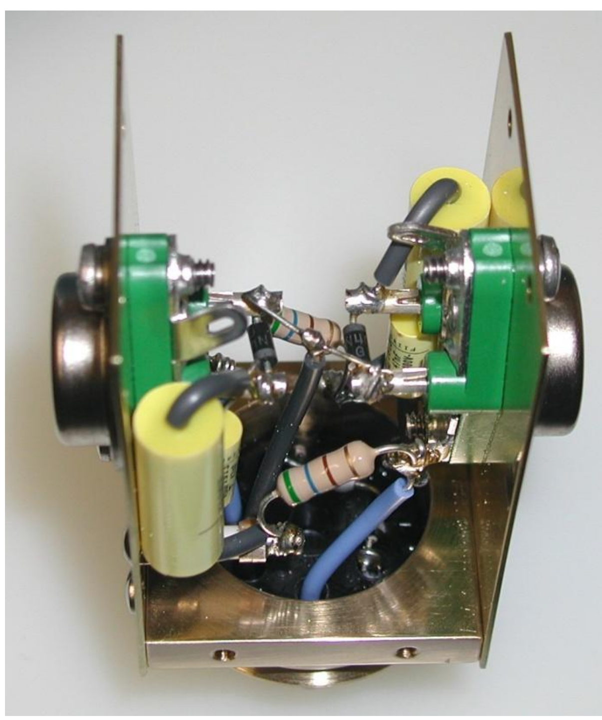
Each transistor Base has two capacitors & one diode connected to it. There are no resistors connected to the Bases because the base resistor is internal to the Darlington transistor itself. The scribe marks for the holes in the brass plate are on the inside surfaces so they are not visible from the outside of the assembled unit.

The brass plates have also been sprayed with DS117 clear automotive lacquer to prevent oxidation.

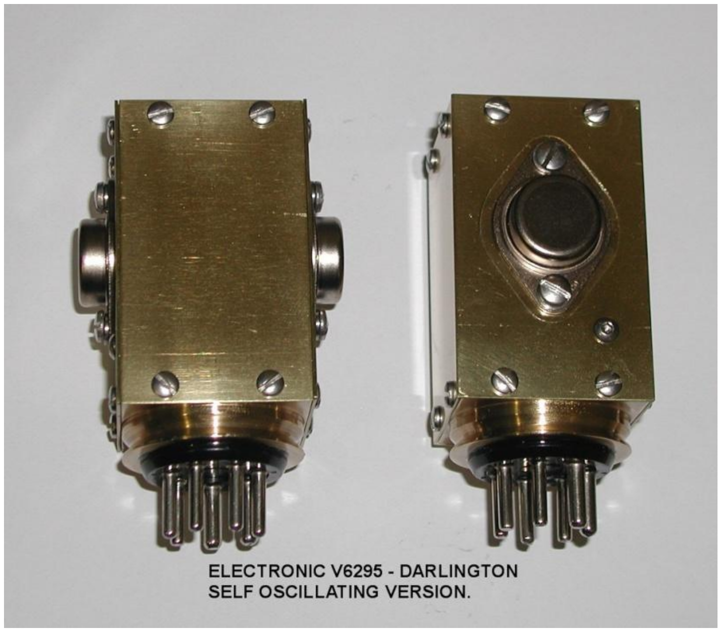
The transistors are mounted with the usual mica insulating washer. Put silicone grease on both sides of the washer. Clear silicone is less messy than the white compound and the extra is easily wiped away. In this instance each transistor’s dissipation is only 1 to 1.5 watts maximum so they only run warm, still it is better to have some thermal coupling to the brass plate:



The transistors are screwed down with ½ inch long 6-32 screws that fit the threads in the Augat sockets. Each screw has a split spring lock washer under its head. The photo below shows the internals assembled.

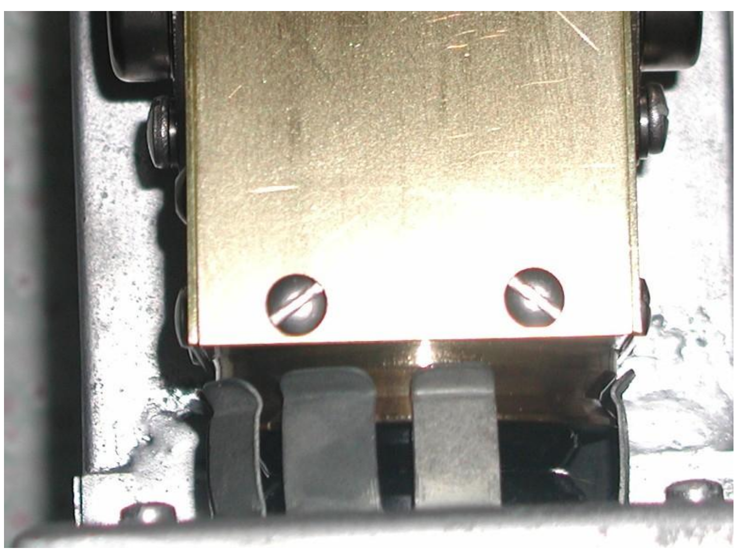


The fixing screws are stainless steel 4-40 UNC, ¼” long with a Binder style head which is similar but a little different to a pan head. These are available from PSME (Precision Scale Model Engineering in the USA).When the unit is assembled, the 560R resistors pass from side to side, connecting the mounting post connection to the collector terminal lug on the opposite transistor.



ELECTRONIC V6295 - DARLINGTON
SELF OSCILLATING VERSION.

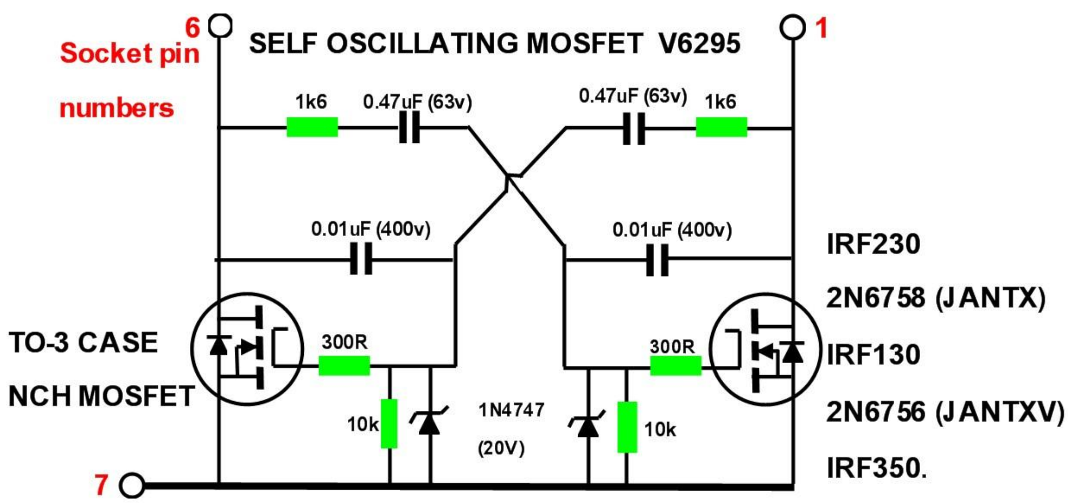
The groove in the base is in exactly the correct position for the clips around the 7 pin vibrator socket in the ZC1:



THE SELF OSCILLATING MOSFET VERSION (added June 2014)

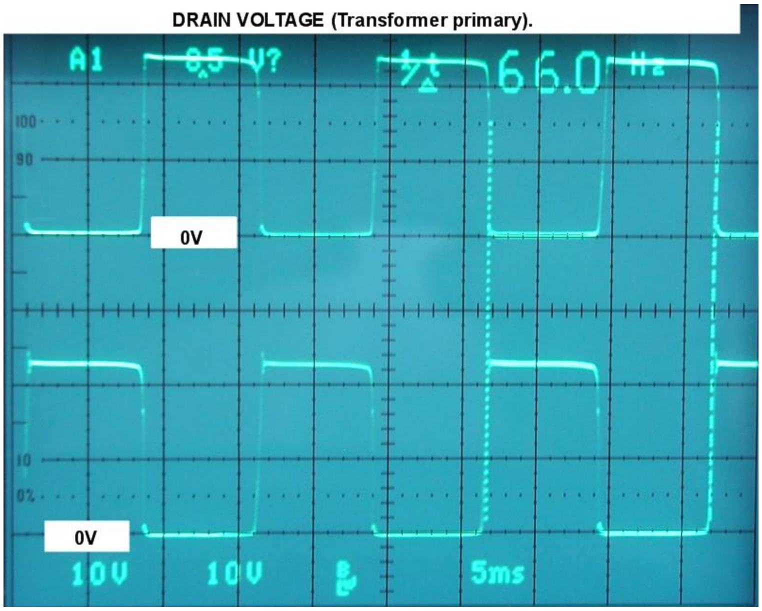
This version has been added because in general the other electronic V6295 versions in cases described above were in a housing which was rectangular rather than round. While this was not a problem for the ZC1 MK2, it could cause a problem in the ZC1 MK1 where the vibrator is mounted close to the edge of the chassis. This version uses a readily available round aluminium housing which is a commercial air intake pipe joiner, 3” long and 38 mm diameter available on Ebay. It also uses a standard Amphenol 7 pin base also usually available on Ebay.

The schematic is shown below:

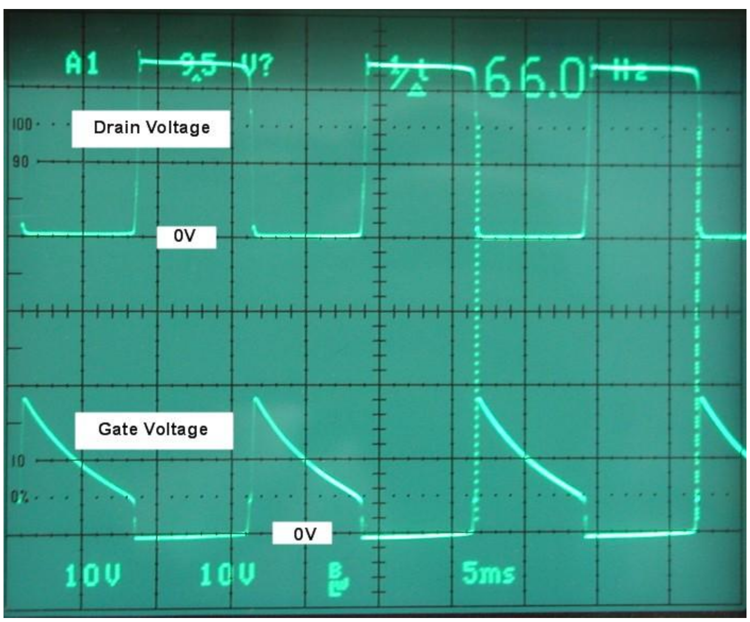


This unit produces very clean switching waveforms and will start from voltages as low as 8V, even when the supply is loaded. Unlike the mosfet unit driven by the independent oscillator it does not require a tuning capacitor on the transformer primary. Also the circuit, much like the Darlington or the transformer/ASZ17 version, is intrinsically short circuit protected because if the supply is overloaded and oscillations stop, both of the mosfets turn off due to the AC coupling. As indicated a number of Mosfets will work in this circuit. TO-3 case versions were chosen, however TO-220 case versions could be used such as the IRF540N.The better quality parts are the mil spec JANTX types.The IRF230 and IRF350 are available from RS components or Element 14. The 2N6758 is available on Ebay, the ones made by Harris are particularly good quality. The IRF350 is a little expensive, the IRF230 is better value.

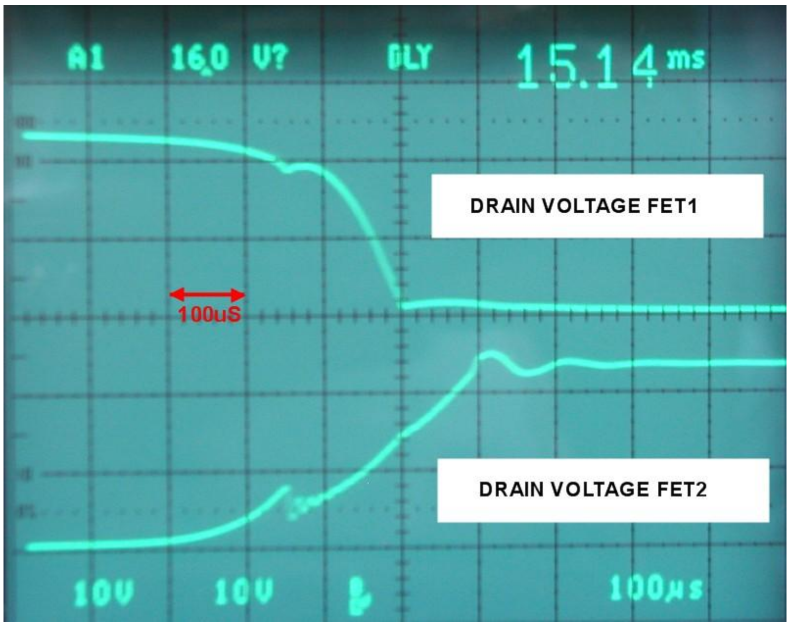
The oscilloscope trace below shows the drain voltages with the unit in operation in the ZC1 in receive mode. These are very similar to the waveforms seen with the Darlington version:



The trace below shows the gate voltage of one of the mosfets when it is conducting. The 0.47uF charges via the drain voltage (24v) of the fellow mosfet and 10k gate resistor until the charge current drops off and the gate voltage approaches the threshold mosfet's voltage. Also by that time there is some transformer core saturation beginning and so the feedback rapidly falls away, the mosfet turns off and the fellow mosfet is driven into conduction. The unit runs at 66Hz:



The discharge path way for the 0.47uF capacitors is more rapid via the 1k6 and the gate protection Zener diodes conducting in their forward direction, so the gate voltage cannot fall below 0.7V. Looking closely at the switching transitions on the transformer primary (drain connections) 10V/div, they are free from radio frequencies and any excessive voltage overshoot with this circuit:

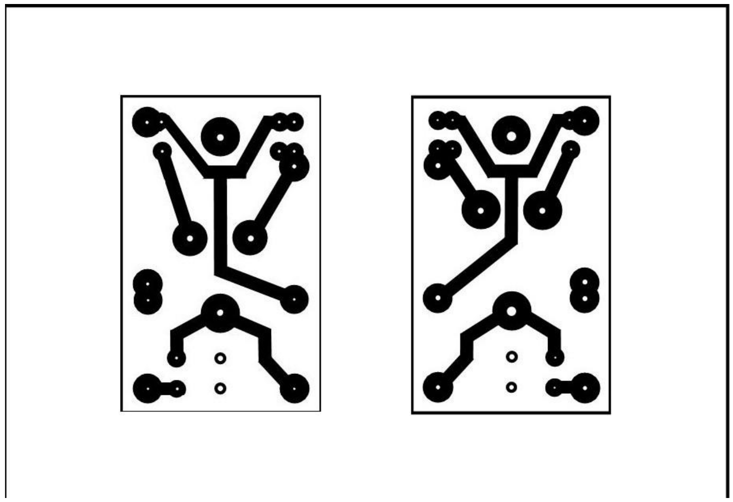


Due to the lower drive requirements of the Mosfet vs the Darlington, the coupling capacitor values can be smaller (0.47uF vs 4.7uF). This assists in making a physical unit which will slide easily inside the pre-made 38mm aluminium tube.

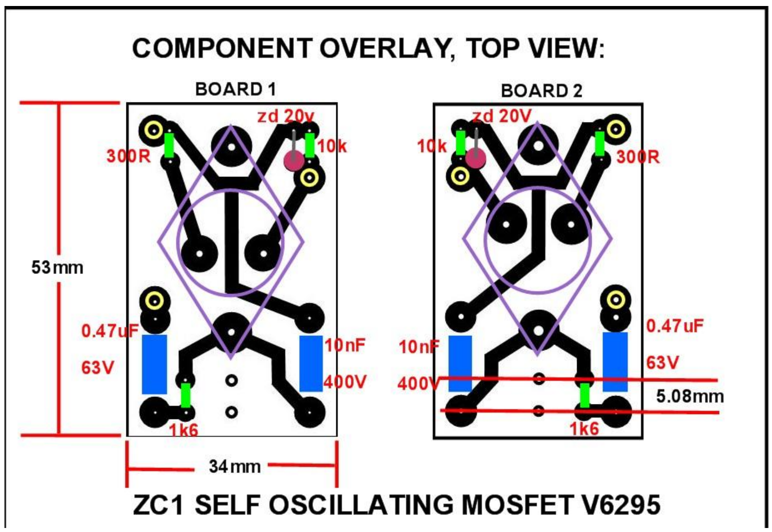
Building the self oscillating V6295:

In this design there is less dependence on complex metalwork. Merely two small pcb's which face each other.

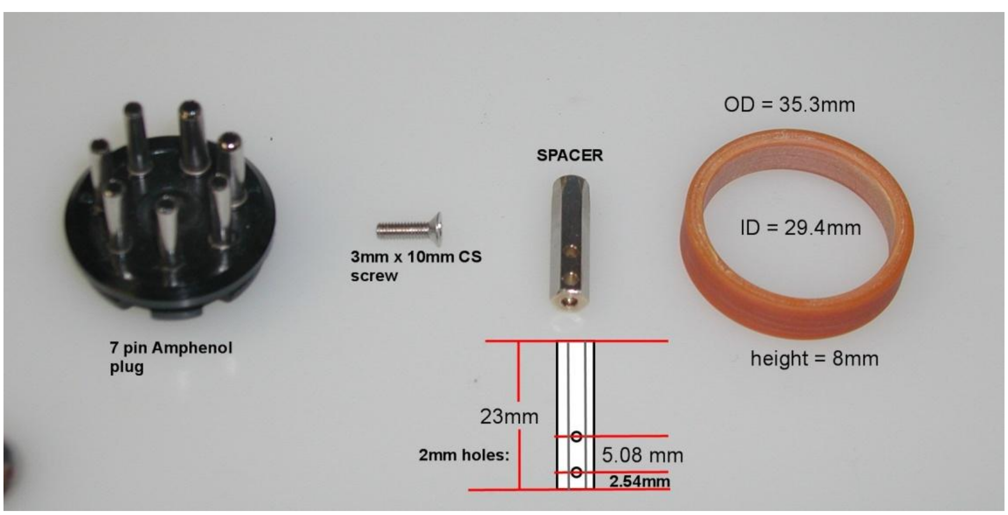
The diagrams below show the two simple pcb's used. The hole spacing for the components all fall on a 2.54mm grid, except for the TO-3 transistor holes which do not land exactly on the grid due to the geometry of a TO-3 package.



The diagram below viewed through the component side of each pcb:

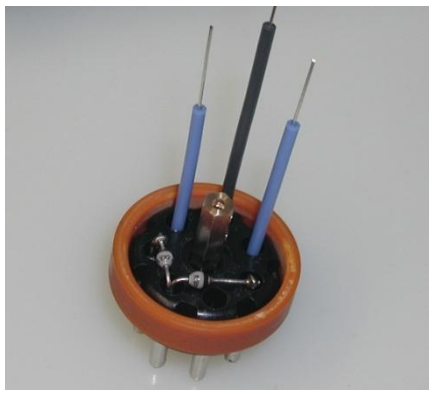


The base is prepared in the usual way with the BY448 diodes and in this case a 5.4mm AF 23mm long 3mm threaded spacer with two 2.0 mm drilled mounting holes:

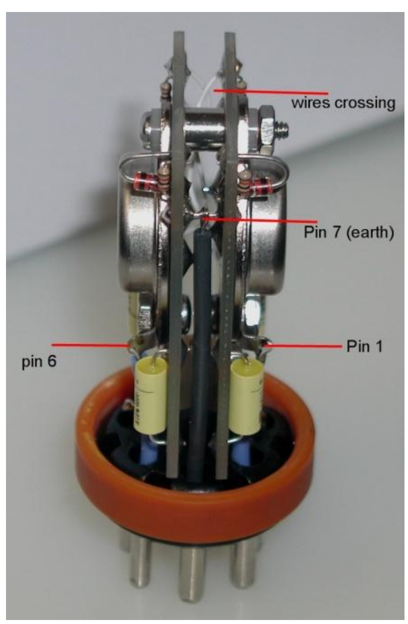


The BY448 diodes are not necessary if the unit is only to be used in a ZC1 MK 1, however it is better to have them so that the unit will work in either the MK1 or MK2.

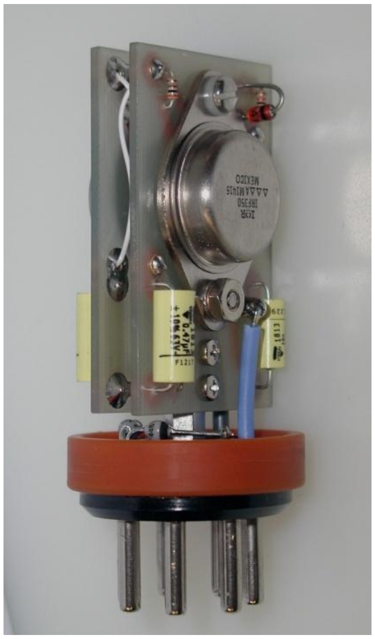
A spacer is required to help fit the finished unit into the aluminium tube. This spacer is 35.3mm OD, 29.4mm ID and 8 to 10mm tall is satisfactory. The one shown in the photo above was cut out of a piece of phenolic plate with two hole saws then trimmed to size. This spacer can also be made of metal like aluminium for example. The photo below shows the assembled base. The spacer is glued to the Amphenol base with 24 Hr Araldite:

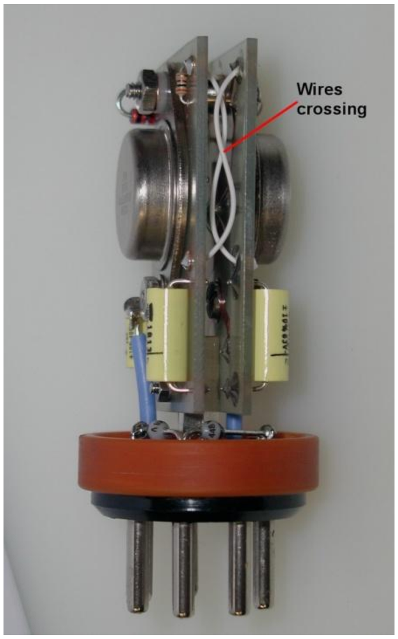


The two pcb's face each other with a 5.4mm gap between them. The earth connections is arranged so a small length of 0.7mm tinned copper wire passes between the pcb's. Two other "crossing wires are required. These are best illustrated by the photos:



The photos below show the two crossing wires. These are Teflon covered wire wrap types however any small hook up wire would work:

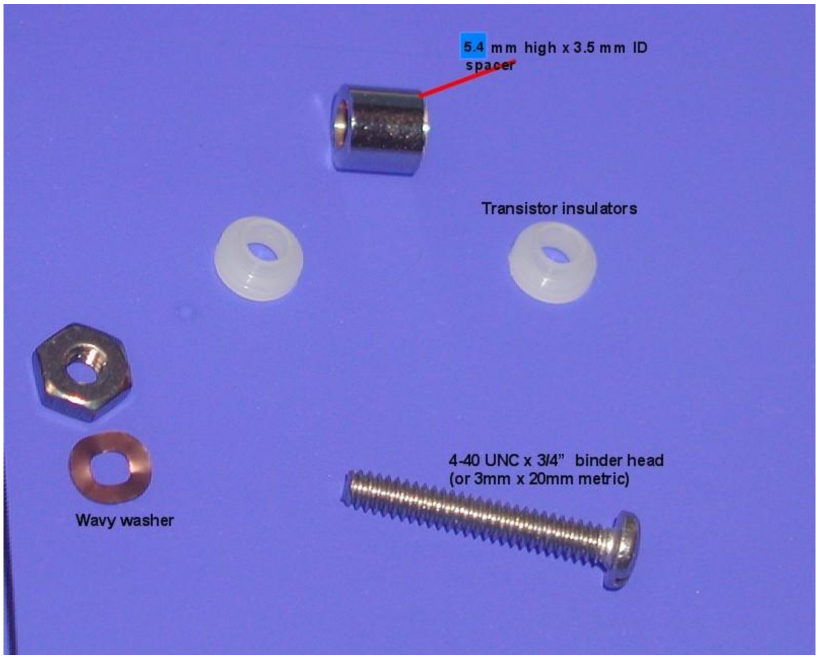




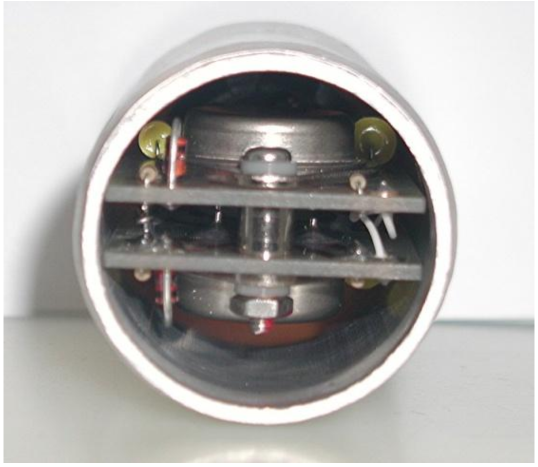
Ideally the pcb has plated through holes. In the absence of those in this hand made prototype small brass eyelets were used for the connections between the pcb's)

The lower TO-3 transistor mounting hole has a screw & nut to secure a lug for the Drain connections to pin 6 and pin 1 of the amphenol base. The pcb assembly is rotated so that the gap between the pcb is over pin 7.

This allows the wires from the base to pass in a very direct & orderly way to the earth and two drain connections. The upper mounting holes between the transistors are secured with a spacer & some insulators:



The final procedure is to fit the unit into the pre made aluminium tube. The top of the tube is sealed with a 35.3 mm diameter 6mm thick disc glued into place. This is a firm press fit. The disc shown in the photos was made from Bramite which is a fibreglass like insulator, however it could be made from aluminium or Paxolin or any other material. The photo shows a view looking into the unit before the top disc was fitted:



The base is a firm fit into the tube. One option would be simply to glue it in because it is unlikely this unit, due to its rugged nature, would ever require repairs. However it turned out that it could simply be retained with a 1.2mm to 1.4 mm spring clip made from spring steel wire. This engages the existing groove in the aluminium housing and then some varnish is applied. This allows disassembly if required one day:

The photo shows this unit next to the original V6295. The exhaust pipe tube is about 1.5mm greater in length than the original housing, so I trimmed 1.5 mm off the lower edge (at the base end), but this is not absolutely necessary, it still fits well in the socket without that done:

Original V6295 **Self Oscillating Mosfet V6295**

V6295 Efficiency Shootout :

1) Self Oscillating unit with Transformer coupling & ASZ17 transistors (ASZ17).

2) Mosfet V6295 with internal oscillator (MOSFET).

3) Self oscillating unit without transformer & Darlington Transistors (DARL).

4) Original Electromechanical V6295 (MECH).

5) Self Oscillating mostfet version added June 2014.

The table below shows the results obtained with the ZC1 power pack and a dummy load on its output.

ZC1 MK2 POWER PACK RUNNING VARIOUS UNITS:
12V SUPPLY and 3750R Load + 47uF capacitor.

	DC OUTPUT VOLTAGE	EFFICIENCY % $\frac{\text{POWER OUT} \times 100}{\text{POWER IN}}$
MECH	267	66.6
DARL	265.4	67.3
MOSFET	289	72.7
ASZ17	282	67.5
SELF OSC MOSFET	276	67

(All the electronic versions use BY448's pairs to replace the secondary contacts)

The most efficient unit with the highest output voltage too, is the oscillator driven mosfet version. This is to be expected because of the low power drive requirements for the mosfet's gates and the mosfet's (IRF540N) low RDS on. The Self oscillating version uses mosfets with a slightly higher RDS on accounting for the differences there.

What is of note is how surprisingly good the electromechanical unit is. (This unit is in top shape and properly adjusted). It just goes to illustrate an interesting point: Never underestimate the genius of the designers of electro-mechanical equipment from yesteryear, their designs might surprise you.

The Darlington version is almost a dead ringer in performance to the electromechanical unit, but of course with no reliability or wear issues. The reason the output voltage is a little lower than the other electronic units is due to the collector-emitter voltage drop of about 0.9V for the Darlington. On the other hand the reason the output voltage is lower for the mechanical unit is the reduced duty cycle compared to the electronic units. So the DARL unit and the MECH unit land on about the same performance parameters for two different reasons.

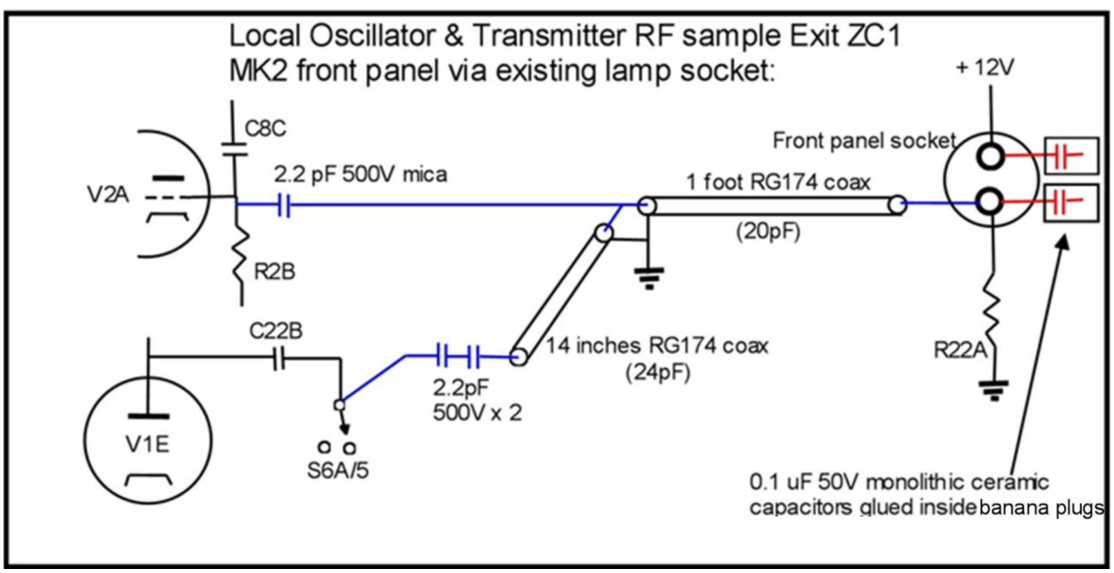
4) ADDING A DIGITAL TUNING FREQUENCY METER TO THE ZC1:

The makers of the ZC1 included a front panel connector which had one terminal connected to the 12V power and the other connected via a 100 Ohm resistor to ground, this is designated R22A on the schematic. This was to power a small lamp for night operations.

A sample of both the radio's local oscillator (L/O) and RF output from the VFO can be applied to this resistor/socket connection to allow these signals to exit the radio for monitoring by a frequency counter.

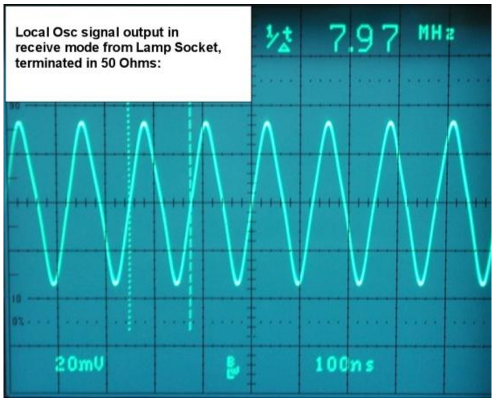
In the receive mode, the frequency counter can be pre programmed to subtract 465KHz (The ZC1's IF frequency) from the displayed frequency to give an accurate readout of the tuned frequency on the dial. Many frequency counter types using LED displays and TTL counter IC's with a pre-loaded data inputs had this facility such as the DFD4 or the TK-1 from Communications Concepts Inc. More modern versions of frequency counters with programmable offsets based on PIC microcontrollers with LCD displays are available on Ebay as kit sets for adding on to communications radios with analog dials.

The following diagram shows how very low value capacitors are added to couple the signals out via the existing lamp socket connections on the front panel:

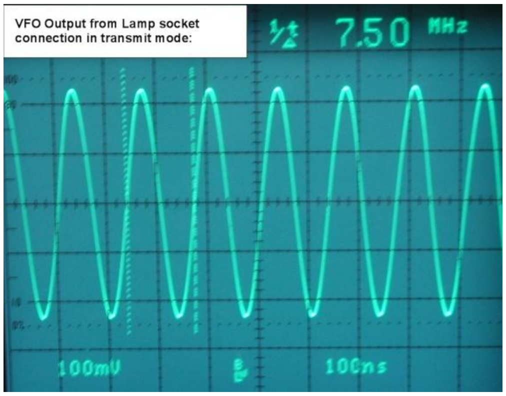


Due to the low values used in the 1 to 2pF range for the coupling capacitors the set barely requires re-tuning after these are added, C7G and C7H trimmers can be adjusted on the transmit side and C7C and C7B on the receive side(L/O) to fractionally reduce their capacity if required but I found it was not necessary. The capacitance of the coax forms an AC voltage divider and impedance transformation. The presence or absence of the external frequency counter results in a negligible effect on receive or transmit frequencies. The use of the original lamp that the front panel socket was intended for, or not, has no discernable effect either. Since one of the connections on the lamp connector circuit is connected to positive and not ground it is a good idea to put two DC isolating capacitors in the banana plugs in case the chassis of the frequency counter and the ZC1 chassis come in contact. In addition the original function of the front panel lamp socket is not altered by this modification.

The following is a recording of the output terminated into a 50 Ohm load in receive mode with the radio tuned to 7.5 Mhz .The local oscillator of course runs 465 KHz (the IF frequency) above the received signal. The peak voltage is only 30mV, not all counters could accept that level and might need a buffer amplifier. My counter has an internal buffer/amp.



In transmit mode the output level is higher at just over 200mV peak. This is useful as the counter can be modified to switch out its 465 KHz offset in transmit mode show it will automatically show the correct receive and transmit frequencies without having to manually switch the offset on the frequency counter.



The frequency counter is plugged into the lamp socket as shown. This particular counter is basically a Communications Concepts Inc type TK-1. I added an extra two digits for a 1MHz and 10 MHz column and the displays were changed from the usual 7 segment to TIL311 types. The TK-1 was very useful as it has internal DIP switches so that the counter can have any IF frequency offset subtracted from the measured frequency. This enables it to display the tuned frequency in any radio where the local oscillator runs the IF frequency above the received station, which is the usual case.

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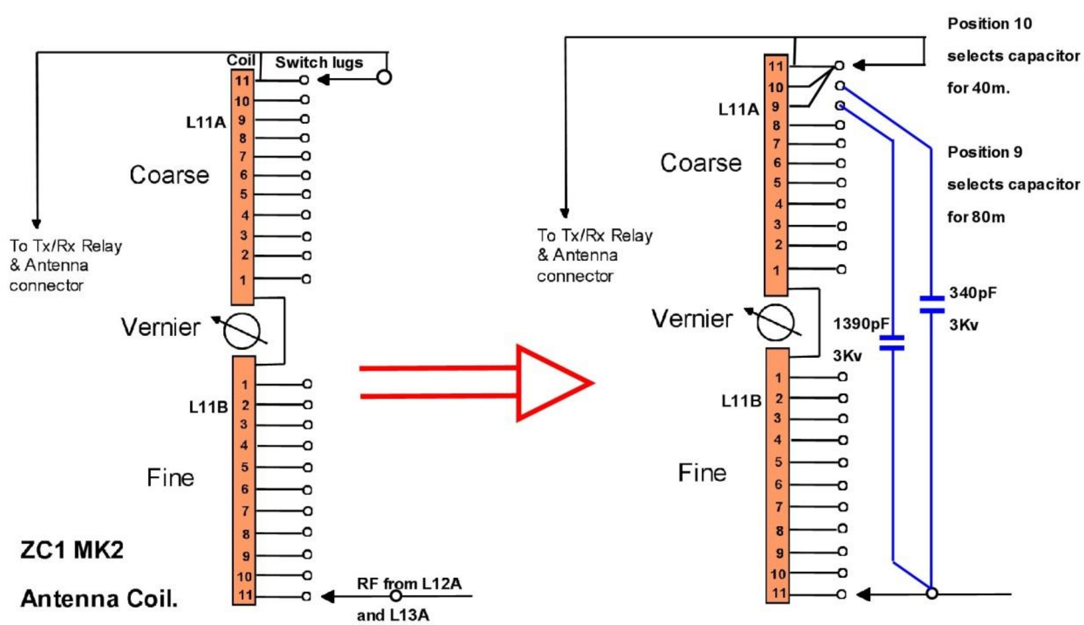
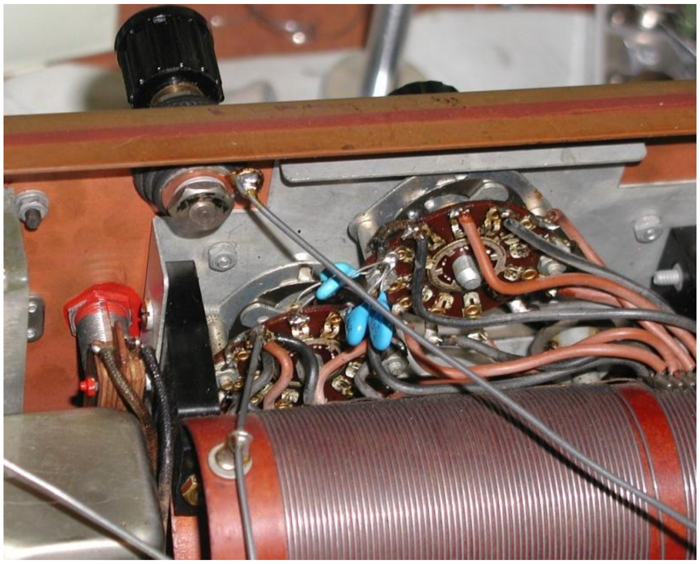
5) OBTAINING MAXIMUM RF POWER INTO A 50 OHM LOAD ON 40m & 80m.

As the reader knows the ZC1 MK2 has a very impressive loading coil to allow an external antenna to be matched. Experiments verified that the direct outputs of the Tank coils L12A and L13A (prior to the loading coil) correctly capacitively coupled into a 12.5 Ohm load produce the maximum output power. A 1:2 turns ratio Toroidal transformer transforms this to 50 Ohms. The capacitor values are different for 40 and 80 metres.

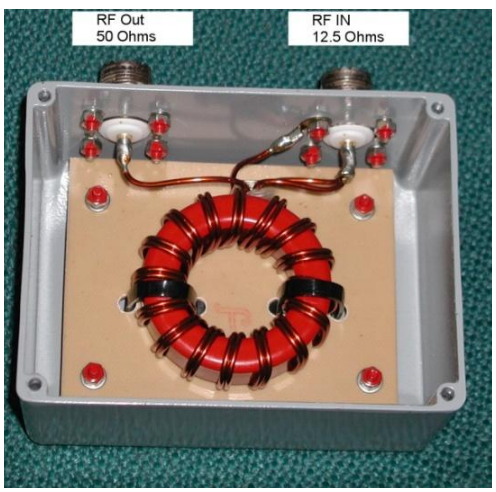
The required coupling values are 340pF for 40 meters (7.5 MHz) and 1390pF for 80 metres (3.75 MHz). The 340pF is composed from a 220pF and a 120pF 3Kv Ceramic capacitors and the 1390pF from a 1000 & 390 pf 3Kv Ceramics. This capacitance is required to cancel the reactance of the antenna coupling coils of L12A and L13A. The values chosen peak the power output on 80m and 40m.The first step is to add these to the fine & course antenna tuning switches as shown in the diagram and photo below:



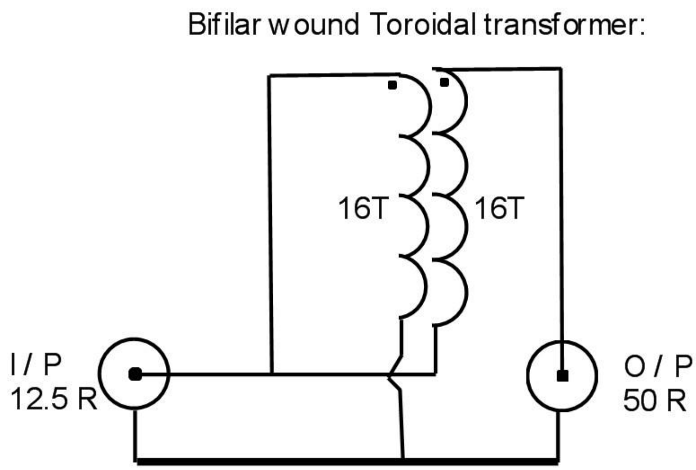
The wires from L11A which are unsoldered from the switch lugs corresponding to position 9 and 10 are simply soldered on to the lug for position 11 so that they are not dangling around. Then all that is required is to select the correct capacitor for the 40 and 80 meter band using the coarse position switch. This modification is completely reversible without any damage to the unit. The photo below shows these capacitors sitting on the rear of the switch unit:



The antenna output connector is then coupled with 50 Ohm coax into a box containing a 1:2 turns ratio (1:4 impedance ratio) matching transformer which is created from an Amidon T-200-2 toroidal core kit (part AB200-10). The transformer is bifilar wound with 16 turns of 1.7mm diameter enamelled copper wire (which comes in this Amidon kit) and connected as a 1:2 autotransformer. It can be mounted on a rectangle of blank pcb (all copper removed) and mounted on spacers inside a diecast box with the UHF connectors. If it is a painted box like the one shown make sure to clean the paint away from the connector mounting screw areas so bot the connectors are thoroughly earthed to the box on each of their 4 screw points.



Electrically the transformer is configured as shown:



The Amidon kit comes as follows:

MEASURED POWER OUTPUT WITH THE AMIDON TRANSFORMER:

The measured power output into a 50 ohm dummy load is 3.26 Watts on 80m and 2 watts on 40m.

It is possible to increase the power beyond this with circuit modifications however the simple system described above gets the most out of the standard ZC1 unit into a 50 Ohm load.

The photo below shows the unit under test with the impedance matching transformer and 50 Ohm power dummy load:

If 100% modulation were possible the peak envelope power (PEP) on 80m would be 13 watts. In practice the amplitude modulator in the ZC1 is not able to achieve this and in addition the microphone signal is not high enough in level to fully deploy the modulator either. However with an external audio source, instead of the microphone, I have measured a peak envelope power PEP of 9 watts on 80m and 7 watts on 40m. Which is quite impressive for the pair of 6V6's composing the RF output amplifier and the Amplitude Modulator stage.

Note: With the capacitors described above alone and no impedance matching transformer used; the power output is 1.7 watts on 80m and 1.1 watts on 40m into the 50R load. The transformer results in doubling of the output power so it is well worth the work to make it.

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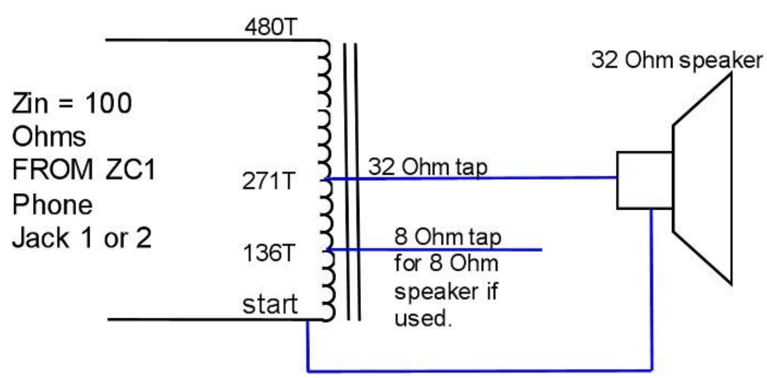
6) EXTENSION SPEAKER REPLACES HEADPHONES:

The ZC1 uses a 6U7-G radio frequency tube (triode connected) as the audio output amplifier. Testing this circuit shows that designers pushed this tube to near its maximum ratings, which is a plate dissipation of 2.25 max watts and a screen dissipation of 0.25 watts, or a total tube dissipation of about 2.5 watts. The available audio output power prior to clipping is 6v peak into 100 Ohms or about 180mW, however, the distortion increases over 100mW output. The 2K cathode resistor for the 6U7 can be reduced to 1k8 to gain a little more power, which is in the range for the specification of the original carbon resistor, most have gone high in resistance to the range of 2.5k to 3k. In addition if the tube is switched for a 6K7, which has higher plate dissipation, but otherwise similar to a 6U7, the cathode resistor can be lowered to 1k2 which gives a good improvement.

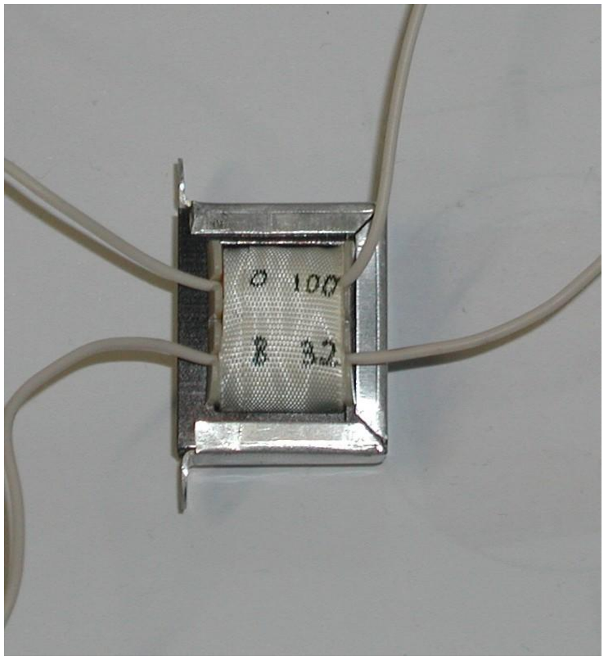
Keeping the set original and trying a 32 ohm extension speaker gave a surprisingly good result. However, the impedance mismatch increased the distortion for any given power level. It is best to match the speaker with a small autotransformer, the design of which is shown below. The taps can be placed to suit any impedance speaker (remembering that the impedance ratio is the square of the turn’s ratio). At this low power level the transformer described here has a flat undistorted response from 50Hz to 20KHz.

Audio Autotransformer for ZC1 External Speaker:

1 cm square inside the bobbin.
Total 480 turns of 0.315mm diameter wire(0.35mm with enamel)
Taps at 136 and 217 turns.
(DC Resistance of 480T = 6 Ohms).



The transformer was placed inside a speaker box, in this case with a spare 32 ohm speaker, however any speaker is ok with the correct transformer tap:



This arrangement gives good results and the ZC1 can fill a small room with sound.

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