400 Hz inverters for air-borne radios. Part 1

Ulrich Fierz HB9AIK/VK2ANI

VMARS members had tried to design 400 Hz inverters based on a recently available printed circuit board (PCB) from China. Even though a newer version of this board uses a SG3525 PWM controller, it produces square waves only, has no regulation and does not make use of the 'modified sinewave' concept. So there were questions raised on the VMARS forum as to how to obtain a 'sinewave' output and what waveform was really needed for aircraft radio power supplies. This article aims to clarify some issues and provides tips on how to proceed. It will consist of two parts: Part 1 deals with the background, specifications and existing solutions and problem areas for amateur construction analysed; a breadboard layout using the Chinese PCB in question and test results are presented. Part 2 will describe an up-to-date design and implementation of a 400 Hz inverter so that it can be built/copied using available parts, again comprising a PCB from China.

Background

A standard for larger aircraft electrical power supplies has been developed over the years and is basically still used today. It consists of two sources, a 400 Hz three-phase 115 V AC generator and an alternator feeding 28 VDC to a battery – in the past, this battery would have been a substantial NiCd unit.

The uses for the two sources of power can roughly be summarized as follows:

The 400Hz three-phase supply is used for powerful motors, heating the galley, air conditioning, *etc*. as well as some of the navigational aids and, sometimes, unfortunately for us amateurs, to power radios with high power output. A common example is the Collins 618T-2 capable of 400 W p.e.p. output. There is no practical way to generate this kind of three-phase high power supply except using one of the very heavy, large and expensive frequency converters built for this purpose.

The DC power supply is used to run all kinds of devices such as navigational aids, computers, cockpit indicators, radios that are, for example, needed even when the engines should fail. An example of a radio to be used in such a situation is the Collins 618M VHF radio series with some 25 W output.

A 400 Hz three-phase supply is, however, usually not provided in small to medium-sized aircraft or helicopters and everything is run from the 28 VDC system; 14 VDC is sometimes used in small planes.

Nevertheless, many navigational aids and radios, still required a single phase 400 Hz 115 V supply mostly for servo control/tuning systems, which now needs to be produced somehow from the DC source; a 400 Hz DC/AC inverter was required. In the early days, this AC supply was generated using a dynamotor, for example, a 400Hz output was integrated into the main dynamotor in the Collins 416W-1 power supply for the 618S-1 radio. Eventually dynamotors were replaced by solid-state circuits. The specifications of some solid-state inverters used in aircraft are given in **Table 1.**

This article will describe only solid-state implementation.

Parameter	AI^2 1A250	Collins 488A-2	Collins 426T-1	C.F.E. 200810
Input voltage	28 VDC ±15%	27.5 VDC ±10%	27.5 VDC ±10%	27.5 VDC ±15%
Waveform	sine-wave	sine-wave	modified sine-wave	modified sine-wave
Output voltage	115 VAC ±5%	115 VAC ±10%	120 VAC ±10%	120 VAC ±5%
Output power	250 VA	250 VA	630 VA	700 VA
Power factor	0.8 lag – 0.9 lead	0.8–1.0 lag	–	–
Frequency	400 Hz ±1%	400 Hz ±5%	400 Hz ±1.5%	400 Hz ±1%
Distortion	3% max	10% max	–	–
Efficiency	75% typically	65% min	75% min	75% min
Weight	9.3 lbs	8.4 lbs	16 lbs	16 lbs
Year	1973	1963	1962	1983
Status	in regular use	retired	retired	in use

Table 1. Specifications of some US-made inverters used to power aircraft equipment [1]

Commercial solid-state inverters

Before considering amateur projects, let us take a look at how solid-state commercial avionics units evolved:

Classic solid-state inverters use a switching technology to generate directly a 400 Hz rectangular pulse-width controlled waveform, now often referred to as a 'modified sinewave'. For many 150–250 VA units, the inverter output was passed through filter configurations to approach a true sinewave. Developments in this technology can be seen mostly in the control circuitry, which moved from unijunction transistors and saturating inductors to discrete logic circuits with flip-flops and dividers, and replacing germanium with faster silicon transistors. Eventually, the control circuitry used integrated circuits (ICs) such as pulse-width modulation (PWM) chips as drivers and operational amplifiers or specialized ICs in the control loop.

In order for inverters to work satisfactorily at the desired output frequency, all transformers, chokes and other critical components should be designed for operation at 400 Hz. As can be noted from the weights of the inverters given in **Table 1**, there is nevertheless a substantial amount of iron in these units but it allowed germanium power transistors with a very low f_t and very low V_{CE} to be used successfully. The resulting AC waveform out of filtered units with less than 10% distortion appears good on an oscilloscope but, as will be shown below, is not easy to obtain.

Modern 'true sinewave' inverters use a different approach to generate, or rather synthesise, a sinewave. The PWM chip, or a suitable discrete circuit, works at a higher frequency, *e.g*. 12–20 kHz, than the desired output and the width of the pulses is modulated with a sinewave. This results in a pulse train of variable width and, after low-pass filtering carried out at a high cut-off frequency (thereby saving on large inductors and capacitors), the outcome is a rippled sinewave with much less than 10% distortion. An extensive description of this system is found, for example, in **[2]**. Suitably programmed microcontrollers are now often used to generate the required waveforms.

The so called 'true sinewave' inverters, currently available on the market for 50/60 Hz supplies, make use of these principles. They do not use a transformer at the output frequency but there is a 'primary' inverter to generate a high DC voltage corresponding to the required peak-to-peak voltage swing, which is then switched according to the PWM information from the controller and subsequently filtered.

A PWM inverter for 400Hz, without using a primary inverter but a 400 Hz transformer, and its construction for the use in an amateur station, will be described in Part 2.

Required performance

Logically, the required performance characteristics of such an inverter depend on the intended use.

The 1A250 inverter (**Table 1**) with the highest 'quality' output, made by Avionics Instruments Inc., was designed to supply all installed avionics equipment needing 400 Hz 115 VAC or 26.5 VAC, usually in a dual redundant configuration on smaller aircraft **[3]**.

While a power supply for navigational aids and other critical items needs close tolerances on its output, an inverter such as the Collins 488A-2 (**Table 1**) designed for the 618T-3 can have a more relaxed output specification. Similarly, the Collins 426T-1 was used, for example, in the MRC-95 for powering the 618T-3 radio, 490B-1 tuner and a Kleinschmitt teletype. This works well as long as there is a sufficient base load.

Amateur radio projects

Evolution of projects

Early on in the development of inverters, it was found that simple square wave inverters sold to consumers would not work as a power source for our radios and tuners. The square waveform does not provide the peak voltage of a sinewave (1.4 x RMS voltage) which is needed in a power supply with a capacitor-input filter. Also many phase shift networks or motors, for example in servo applications, do not perform as intended. However, such inverters were suitable to operate electric hand tools or kitchen appliances while camping.

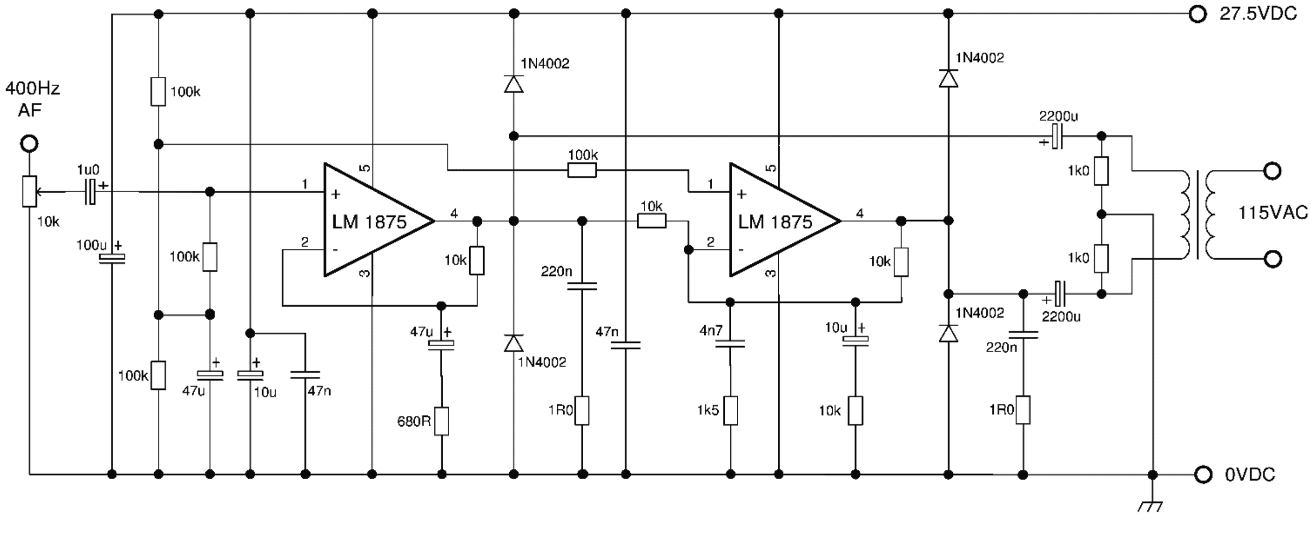

Figure 1. Circuit diagram of inverter based on an amplifier currently-available audio chip

Figure 2. IPS003 power stages in a three-phase configuration for navigational instruments

In 1995, the author required a >15 VA 400 Hz sinewave power source for his 180L-2/3 tuner; a suitable inverter using a pair of TDA2030 class A integrated audio amplifier chips to provide up to 50 VA of true sinewave power was built from a parts kit **[4]** and tested. **Figure 2** shows a module (IPS003) with the solid-state circuitry of three such inverters on one board to provide a three-phase supply for navigational instruments. This type of inverter circuit can still be built today with LM1875 chips (**Figure 1**) remembering to add a large heat sink and accepting poor efficiency

As a next step, 'modified sinewave' inverters, while not a solution for everything, perform at a much higher efficiency than a class A amplifier. Such inverters are sold widely for 50/60 Hz applications and there are several projects to modify these inverters for use at 400 Hz to provide aircraft radio power for the amateur. Each model is different and PCBs with surface mount devices are not always easy to analyse and modify.

In modified sinewave inverters, the duty cycle of each half-wave pulse is reduced and the peak voltage increased to approach the wanted peak value. The period (one cycle) of a 400 Hz waveform is 2.5 ms. If, instead of square wave pulses of 1.25 ms duration, each half-cycle consists of 850 µs pulses, a sinewave with distortion of just less than 30% results (no filters used). Voltage regulation is possible by varying the pulse width, but at the cost of increased distortion. High distortion, *i.e.* higher harmonic content of the waveform, can reduce the performance of electronic equipment powered by the inverter. If voltage regulation is implemented, the arrangement must keep the pulse height constant and the true RMS value controlled. To do this properly, the AI² 200810 inverter actually computes the true RMS value with a complex AD536 chip.

Recent experiments and descriptions by other amateurs confirmed that modified sinewave inverters can be used successfully, *e.g.* with a 618T-3 radio and 490T-1 tuner, a solution comparable to using the Collins 426T-1.

A 'modified sinewave inverter' is an attractive solution with a PWM chip as driver and modern high-voltage MOSFET power transistors, offering good performance at low cost and wide availability. Commercial units use a two stage concept: first an inverter at, for example 100 kHz, converts the input DC to a DC voltage corresponding to the required peak voltage (with or without galvanic separation). The second stage then generates the modified sinewave directly from this high voltage, often by means of a bridge circuit. No transformers at 400 Hz are used.

However, an amateur solution would avoid high voltages and consequent risk of RF interference by using a 400 Hz transformer as in the older systems. However, the power transformer, which should preferably be a toroidal unit of sufficient VA-rating, is likely to be costly.

Adding filters

Starting with the description of switched inverters above, there remains the question as to how a better waveform, conforming closer to the actual shape of a sinewave, could be obtained. Amateurs know well that this is achieved through the use of suitable filters.

To get an idea of what is required to obtain the distortion values in the inverters shown in **Table 1**, their filter circuits were examined as follows:

The 488A-2 uses a series choke followed by a 'box' in parallel (traps) with the output; no further information is available. Perhaps it contains tuned-circuits which are resonant at the 3rd and 5th harmonics.

The 1A250 uses a 400 Hz series-resonant circuit followed by a 400 Hz parallel-resonant circuit plus a series-resonant circuit on 1200 Hz across the line as a trap for the 3rd harmonic (**Figures 3 and 4**).

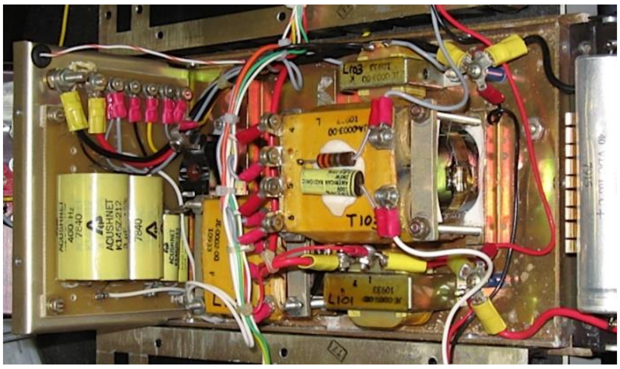

Figure 4. Under-chassis view of the 1A250 inverter showing the filter components

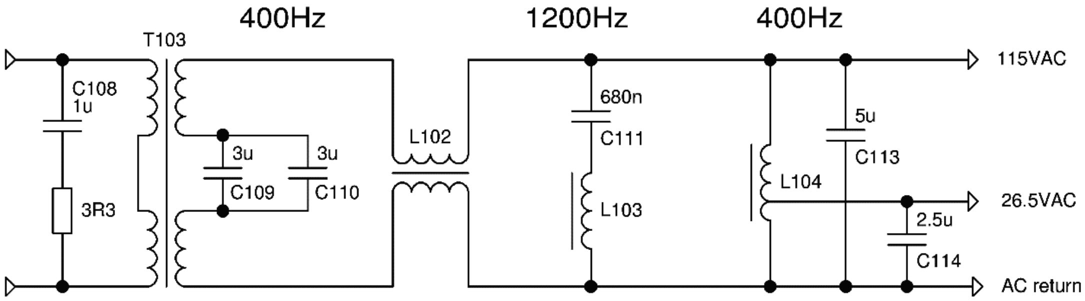

Figure 3. Partial circuit diagram of output filter of the 1A250 inverter

The above two solutions to filter implementation should come as no surprise to the radio amateur but, as will be shown later, a parallel-tuned circuit alone across the output helps to improve performance and regulation.

The author made a number of measurements to assess the suitability of readily available chokes and capacitors in filter circuits. To avoid saturation at the 250 VA power level, inductances with laminated iron cores of sufficient cross sections have to be used. Components for use at 50/60Hz were available but unfortunately no special thinly laminated material nor suitable ferrite cores for audio frequencies.

All measurements were made with true RMS instruments for voltage or current to account for the non-sinusoidal waveforms **[10]**. All currents were measured ‘contactless’ with a clamp-on unit.

Various configurations were tried using a 100VA power transformer, a 200 Ω load and different series and parallel chokes. Capacitors were motor-start units. Only the large iron units functioned, but core losses (not the copper) were high and most cores became quite hot. The parallel reactor fared somewhat better but still became warm. Regulation and efficiency were poor and no viable topology was found to even approach the performance of the 1A250. At least distortion of the waveform was very low; 6-8% was measured under load.

Adding a 1200 Hz trap did not change the results to any extent; even a 0.47 µF X-2 rated capacitor became warm due to the circulating current.

The main lessons learned from these tests were:

- A series-resonant circuit or a series-L apparently cannot be implemented successfully with chokes designed for use at 50/60Hz.
- A parallel-resonant circuit would probably work, even with chokes designed for use at 50/60Hz.
- Traps seem ineffective when using the same chokes.
- Capacitors must be paralleled to share the current.
- Transformers used must have low leakage inductance and low resistance; toroidal transformers are recommended, *e.g.* as for the IPS003 unit.

These considerations entered into the design for the demonstration project described below.

The ‘Chinese inverter’ project

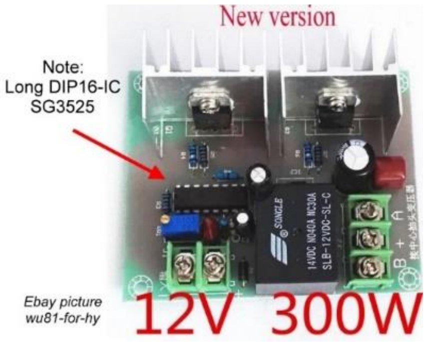

Figure 5. The Chinese inverter module

The inverter unit shown in **Figure 5** is the version used in the current project **[5]**. An earlier product, which is still available, uses a small 8-pin IC that does not provide any PWM functions. The ubiquitous SG3525 chip here is eminently suitable for such an application but unfortunately is not being used to its full potential.

However, it is possible to modify the original circuit to obtain a modified sinewave output and, by adding some additional components, add pulse length adjustment and improve distortion characteristics. Such an inverter, with a modified PCB, has been breadboarded and tested by the author using variable resistive loads up to 240 W dissipation and a 400 Hz 618T fan (about 35 VA at 115 VAC) as a complex load.

The circuit diagram (**Figure 6**) of the inverter shows the installed components as received. 12 VDC input is fed to the board *via* relay K1 which is activated through diode D1 which acts as polarity protection. Power is routed to the transformer terminals and through IC2 to the PWM circuit. A diode (labelled IC2) is installed in place of a 7812 regulator that is needed for operation at input voltages greater than 12 V, *e.g.* 24 VDC operation, and followed by C3. R8 is just a wire. C4 sits across the DC power supply.

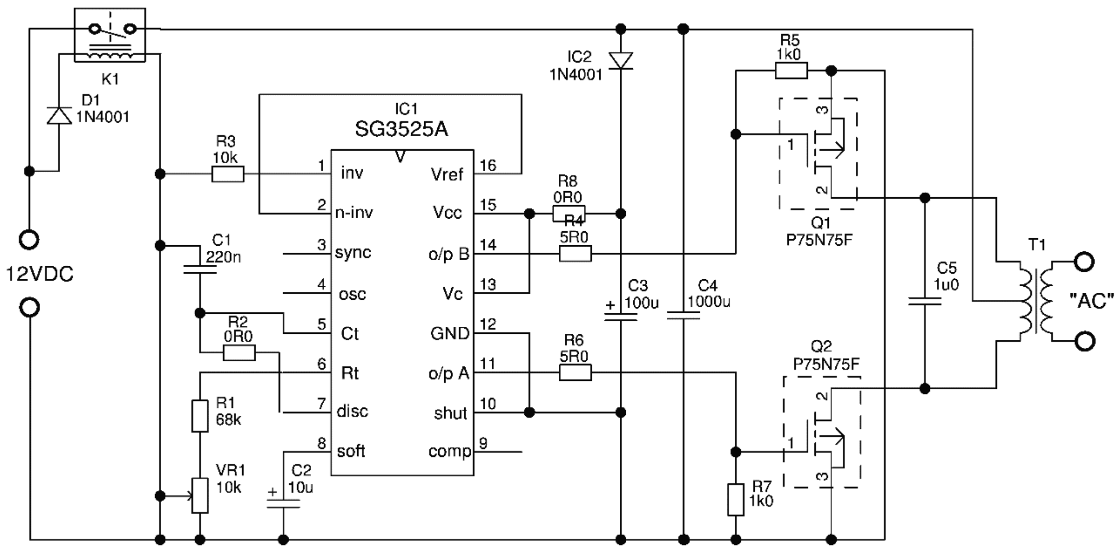

Figure 6. Chinese inverter PCB as received by the author. Copied from the actual PCB by HB9AIK

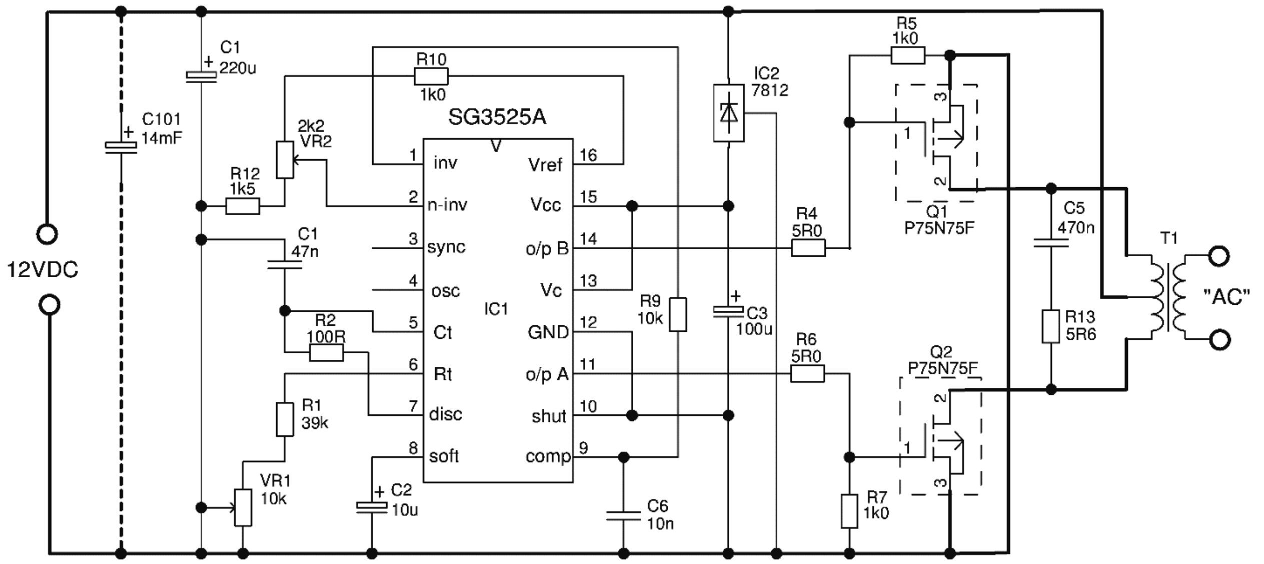

Figure 7. Circuit diagram of the modified inverter PCB

The SG3525 (IC1) feeds the two ST P75N75F MOSFETs Q1 and Q2 through R4, R5 and R6, R7 with pulse signals 180° out of phase. A push-pull transformer is needed to step up the DC supply to the required AC voltage. C5 is supposed to reduce ringing due to the leakage inductance of the transformer. Timing components C1 and R1, VR1 are set to produce 50–60 Hz (which could not quite be achieved in the author's module). C2 provides the soft-start characteristic. The error amplifier is locked to maximum output, the non-inverted input connected to the reference output and the inverting input *via* R3 to ground. The dead-time control is defeated by R2, despite generating a square wave with crossover spikes.

While the basic design is adequate for our purpose, the circuit needed to be modified, as follows (**Figure 7**):

- The timing was set to deliver an output at 400 Hz by changing C1 to 47 nF (film capacitor) and R1 to 39 kΩ.
- Dead time was activated by changing R2 to 100 Ω (could be left defeated in this project).
- A voltage divider consisting of R10, VR2, R12 and fed from Vref was added to the non-inverting input to control pulse width.
- The PCB track from pin 16 to pin 2 was cut.
- 'Comp' was routed through R9, C6 to the inverting input.
- C1 was replaced by a type rated for high ripple current, external C101 added.
- C5 was replaced by a combination of R13 and C5 to improve ringing suppression.
- IC2 may preferably be used if the DC voltage is >15 V (for correct operation, the minimum voltage difference between input and output of the 7812 voltage regulator is 3 V).
- All DC current-carrying tracks were reinforced with thick wire (prior to this modification, the PCB itself heated to over 40°C during the tests).

The modified PCB is shown in **Figure 8**

Figure 8. Layout of the modified PCB

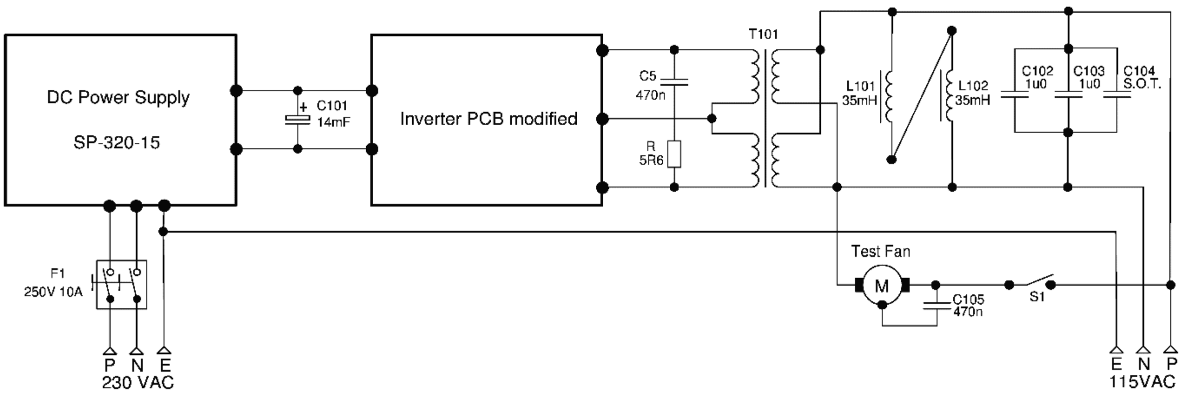

Figure 9. Wiring of the breadboard layout and circuit diagram of the output filter

The circuit thus modified provides a modified sinewave through the ability to set the driver pulse width (VR2) to 850 µs every half cycle at 400 Hz at which setting distortion is at a minimum, providing a good compromise between required peak voltage and RMS power delivered.

A used transformer was chosen for initial tests; at *c.* 130 VA output, the losses (mainly copper, not core) were limiting performance and a 300 VA toroidal transformer was ordered. The results presented below are based on this larger unit.

All measurements were again made using the instruments listed in **[10]**.

Complete setup and results

The breadboard system and its layout are shown in **Figures 9, 10 and 11.** This layout is not intended to be a prototype, although it could form the basis of an implementation; it is more of a demonstration to show how such an inverter would look and how it would perform.

The redesigned power supply provides a filtered modified sinewave output at 400 Hz of up to 250 VA, with less than 20% distortion (including non-resistive loads as exemplified by the fan motor). It has not been made RFI-proof and is without a case, ground plane or RF-filters, so it will most likely produce RF noise. The topic of RF noise will be addressed with the necessary details when implementing the modulated PWM design which will be described in Part 2.

Transformer and DC input voltage

A search for a suitable transformer shows that, apart from making one's own, there is a rather limited selection, especially when looking at toroidal units. EU distributors sell mains transformers rated 200–300 VA with 2x12 VAC and split 115+115 VAC primary windings, with a better choice from the US. In the end a large 300 VA toroidal transformer **[6]** was purchased as the 200 or 250 VA units had too long lead times.

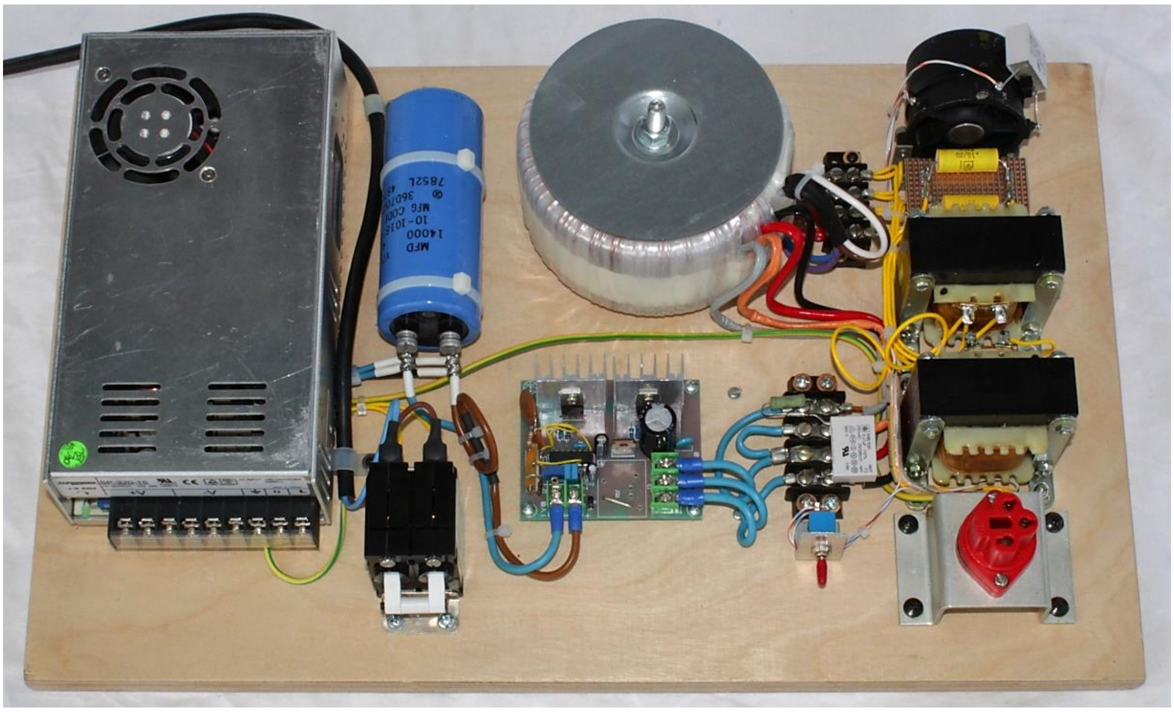

Figure 10. Breadboard implementation of the inverter

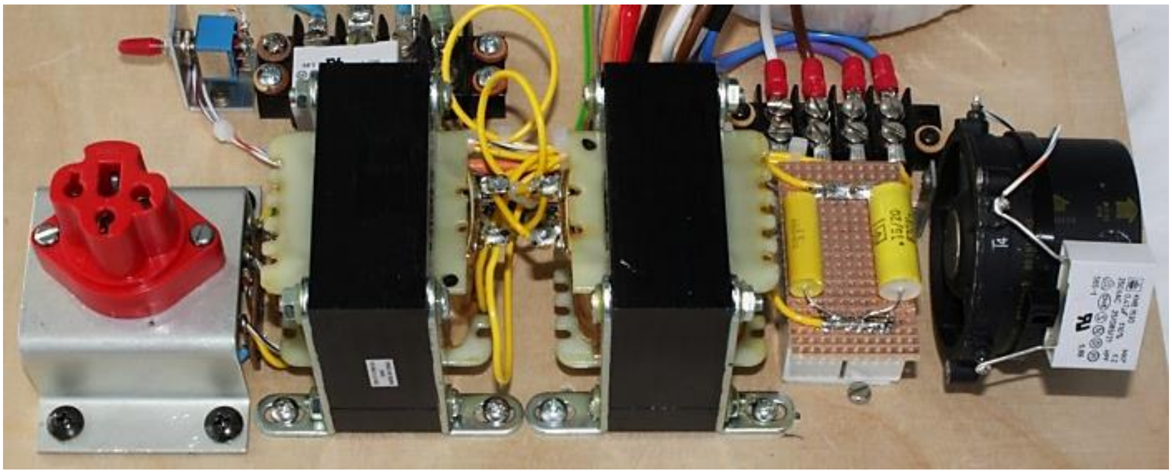

Figure11. Layout of the output filter

It should be noted that the transformer is operated in reverse, *i.e*. the secondary becomes the primary in the inverter and *vice versa*. This has an unwelcome effect as the design of, say, a 115 V to 12 V transformer takes into account the losses at full load, so the unloaded 12 V output voltage will be higher. When used in 'reverse', the unloaded 115 V output will be lower than specified and even lower when under load. For the actual transformer used for this project, the unloaded turns ratio was measured using a low-level 400 Hz signal to be 9.14:1 (12 V+4.8%) compared to the theoretical ratio of 9.58:1 under full load. The reverse loaded ratio, assuming similar losses in the transformer, can then be estimated to be 1:8.7. Thus, the transformer in question becomes 2x12 V to 2x105 V at 300 VA

The next step in the design process is to use these values to determine the required DC voltage to obtain an output of 115 V RMS. The peak value of a pure sine wave is 1.41 x V RMS or 162 V peak. The reader is reminded that a modified sinewave is generated by means of an 850 µs pulse every half cycle, whose duration is longer than the time a sinewave would spend at its peak, so one needs to aim for 140–150 V after transformation. Using the above calculated transformation ratio 1:8.7 this peak voltage requires a DC supply voltage to the inverter of 16–17 VDC. In the breadboard demonstration, this voltage was provided by a 320 W/20 A switch-mode power supply with power factor correction (PFC) **[9]**, set to the required voltage to obtain 115 V RMS under an expected medium load. The estimated voltage requirement was confirmed during tests.

Should a transformer be required to operate the inverter from 13.5 VDC, a toroidal unit that is sold as 115 V:2x10 V at the required AC power could be tried. For operation from 12 VDC a 115 V:2x9 V transformer would prove useful; such 160 VA or 250 VA toroidal units were found available from the US **[8]**. As explained below, the use of 24 VDC or 28 VDC is preferred at this power level.

Output filtering

As outlined above, designing a filter with available off-the-shelf components is difficult. However, some filtering was required to reduce distortion of the sinewave output to an acceptable level. Simplistically, the reason lies in the magnetic energy stored in the large transformer when the MOSFETs turn from ON to OFF and the transformer primary momentarily becomes open circuit. When the transformer is suitably loaded, the load will absorb this energy and the voltage on the secondary falls to zero. With a light load only, however, the voltage remains up at V_{peak} and a square-wave results producing *c*. 150 V RMS on the output.

To manage this effect, a 400 Hz parallel-resonating tuned-circuit was added, similar to the solution in the 1A250 inverter. This parallel circuit acts as an electronic 'flywheel' and the modified sinewave is maintained, and even improved as harmonics are also attenuated, thus lowering distortion to 16–20%. The resonant circuit (**Figure 9**) consists of two 35 mH 2 A-rated iron-cored filter reactors **[7]** sourced from the US in series and paralleled with capacitors amounting to ~2 µF (**Figure 11**). The circulating current in the circuit now is about 1.4–1.6 A RMS and efficiency 75–80% for loads between 30–100% of the rated value. As may be expected, the filter absorbs some energy as loss that is not supplied to the load.

The higher the circulating currents, the greater the effect of the resonant circuit. When using a single 35 mH choke and ~4 µF of capacitance, the circulating current rose to ~2.2 A RMS but efficiency dropped to just over 70%. The core temperature rose to over 55°C, perhaps also caused by the onset of saturation.

The parallel circuit needs to be tuned due to the tolerance of the components. Fed with an LF signal through a 3.3 kΩ resistor from an audio generator, resonance was observed with a voltmeter across the LC-combination and the capacitance adjusted accordingly.

The resulting waveforms from the breadboard layout are shown in **Figures 12–15**, recorded using a digital oscilloscope. As a complex load, the fan does not visibly alter the curves.

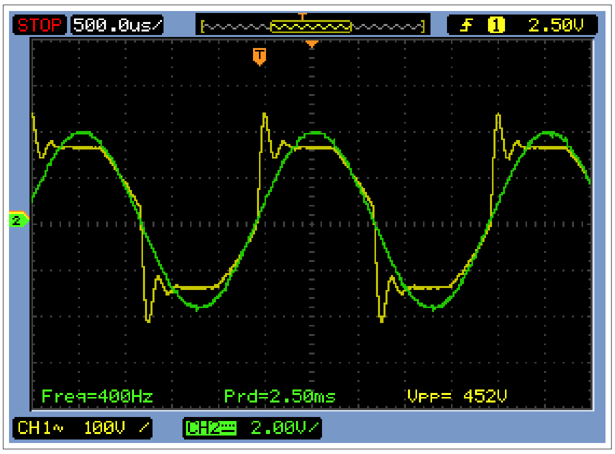

Figure 12. Output voltage waveform with no load (yellow trace). The green trace is a perfect sinewave for comparison

Figure 12 shows the distorted (relative to the sinewave) output when there is a no-load situation. The waveform becomes wider but it still reacts to changes in pulse duration which allows control of the output voltage. The high peak-to-peak voltage is caused by the overshoots at the beginning of the ON phase.

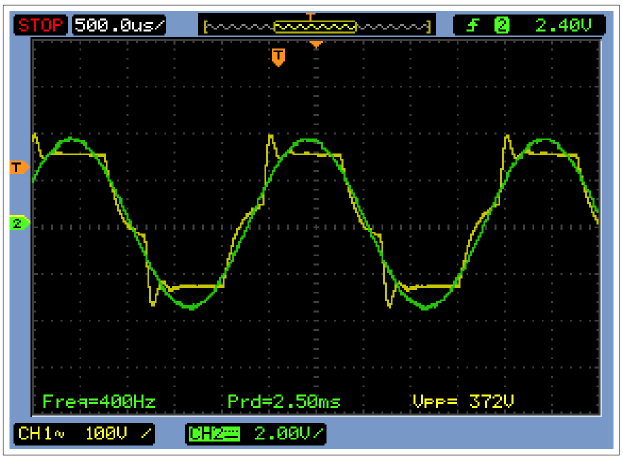

Figure 13. Output voltage waveform at mid-range load (yellow trace). The green trace is a perfect sinewave for comparison

Figure 13 shows the waveform with a mid-range load; here the effect of the parallel-resonant filter has been optimized. This trace also illustrates what may be expected of a modified sinewave.

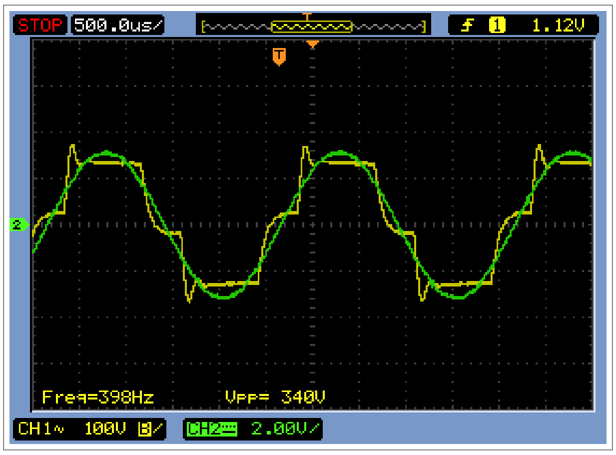

Figure 14. Output voltage waveform at full load (yellow trace). The green trace is a perfect sinewave for comparison

Figure 14 shows the output voltage waveform at full load. The peak-to-peak voltage of 340 V is less than the value of 372 V in **Figure 13** and is caused by transformer losses and lack of stored energy available to the load from the parallel-tuned circuit. There is an increase in distortion, the waveform becomes narrower and has an increased 'step'.

Figure 15 (taken while the output was connected to a mid-range load) shows the peaks and transient oscillations of the input current reflecting a high current peak after turn-on of the MOSFETs. The peak corresponds to the peak in the transformer secondary current in **Figure 16**. This high current peak requires a suitable heavy duty capacitor across the input. The OFF state should be flat; the tilt is caused by the low frequency response of the probe.

Figure 15. Yellow trace: DC current to PCB; Green trace: output voltage. The yellow trace should have a flat response in between peaks

Figures 16 and 17 (taken while the output was connected to a full load) show the equalizing effect of the parallel filter circuit. At turn-on, the transformer current shows a peak which transfers energy to the filter. This turn-on pulse is also present in the filter current where, later at turn-off, the energy is now discharged into the load. As the stored energy is insufficient to maintain the output at full load, the step in the output voltage waveform is created. The traces shown here allow now a better interpretation of the waveforms in **Figures 12–15**.

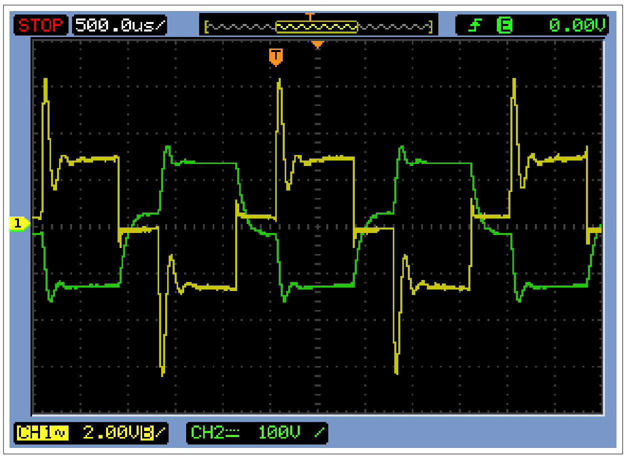

Figure16. Yellow trace: transformer secondary current; Green trace: output voltage

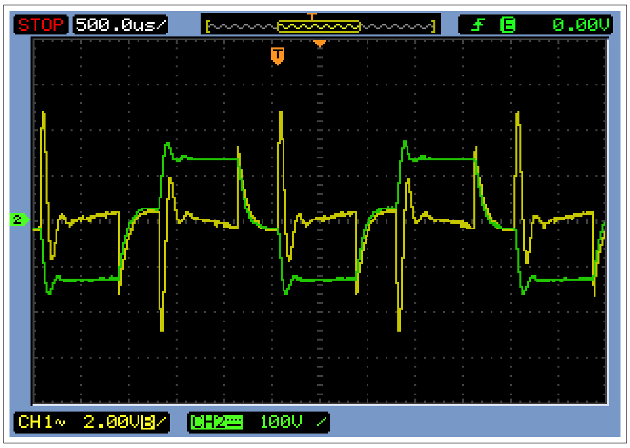

Figure 17. Yellow trace: parallel resonance filter input/output current; Green trace: output voltage

Regulation

The results for unregulated performance into a variable resistive power load are tabulated (**Table 2**). The load being resistive, output power is given in W rather than VA.

The configuration of the inverter undergoing tests did not include any type of voltage regulation. If the load variation is limited and a resulting output voltage spread of ±10% is acceptable, no regulation is needed for avionic radio applications when using a toroidal transformer in the inverter.

For enhanced performance, limited regulation may be effected by reducing the pulse length from the set 850 µs, resulting in limiting the voltage increase at low loads. The AC output would be sampled, either directly *via* a potential divider or a small transformer, the sampled voltage rectified and the smoothed DC signal applied to the error amplifier in the SG3525. Suitable RC networks are needed to set feedback gain and stability. Owing to time restraints, this arrangement was not tested in the current setup where such close regulation is neither advisable nor required by the radios and tuners, but would be more easily achieved in a modulated PWM system.

Conclusion

The experiments described above confirm the difficulty of obtaining a sinewave-like waveform from a simple push-pull inverter circuit, even when using an optimized modified sinewave as a starting point. The heavy filter components required and the increased peak currents in the input and the power transformer require additional expense in quality components that make the simple low-cost inverter circuit unattractive. However, the breadboard system that was designed will successfully operate, for example, a 618T-3 and associated tuner with AC power to spare.

Unfortunately, the Chinese PCB had a number of deficiencies: the board itself, terminal strips, heatsinks and capacitors were all underrated for the currents involved; the project needs to be heavily de-rated or the PCB modified as shown.

The known fact that powers of 250 W or so would be generated better from 24 or 28 VDC supplies was confirmed; losses due to high currents from 12 VDC sources are just too great. The higher DC supply voltage is no problem for aircraft – they run off 27.5 VDC anyway. The design reported here can easily be adapted to 24 or 28 VDC using 115V:2x18 V or 115:2x20V transformers. Even the relay on the PCB is available as a 24 V version.

Part 2 will describe a modulated PWM configuration that will do away with many of the problems encountered here. The author is happy to respond, on the VMARS forum, to any questions relating to the current design.

References and notes

1. The manuals used are: **1A250**: Avionics Instruments Inc., Static Inverters, 1A250, Overhaul Manual, Sept. 1973 (Original). **488A-2**: 488A-2 Static Inverter, Instruction Sheet, 523-0756546-001211, Collins Radio Company, 1 Nov. 1963 (Hardcopy). **426T-1**: 426T-1 Power Inverter, pages 196 – 206, extract from a presumed military manual for the VC-102 system, author and date unknown (copied from VMARS files). **200810**: T.O. 35C1-6-7-1 Technical Manual, 700VA Power Inverter P/N 200810, C.F. Electronics Inc. F04606-82-C-0374, 31 Jan 1983 (Hardcopy).

2. SJ Bitar *et al*. A Pure Sine Wave Inverter. Worcester Polytechnic Institute (US), Major Qualifying Project, 2010-2011, 27/24/2011.

3. The author's 1A250 inverters were apparently used in the PC-6 STOL aircraft. See https://www.pilatus-aircraft.com/en/fly/pc-6.

4. Inverter Power Supply IPS 003, Construction Papers, not formally published, unknown author, ~1992.

5. 300W 12V Inverter Driver, no type number available, Chinese characters only. Search Ebay for "300W inverter driver" and select correct picture with long IC as shown in text. There are several vendors from China, price is less than £10 including shipping.

6. RKD 300/2x12, Block, Germany, from Distrelec.com, p/n 110-80-863.

7. C-56U, Triad Magnetics, US, from Mouser.com, p/n 553-C56U.

8. VTP18-8800 or VPM18-8000, Triad Magnetics, US, from Mouser.com.

9. SP-320-15, Sunpower Technology LLP made by MeanWell Enterprises Co., Taiwan.

10. All measurements were made with an Agilent DOS 1072B digital scope, a refurbished HP 333A distortion analyzer and RMS voltmeter, a Fluke 8024B digital multimeter and a Tenma 72-6185 AC (true RMS)/DC contactless clamp ammeter. Calibration of the HP 333A was checked with a sinewave against two Fluke 8024Bs.

DC-V	DC-A	Pin W	AC-V	∆ V	Distortion	AC-A	Pout W	Efficiency
16.60	2.42	40	132	15%	20.5%	0	0	0%
16.60	4.25	71	130	13%	17.5%	0.30	39	55%
16.60	6.9	115	122	6.1%	16.0%	0.70	85	75%
16.60	9.1	151	118	2.6%	16.0%	1.01	119	79%
16.60	11.7	194	115	0.0%	16.5%	1.35	155	80%
16.58	15.0	249	110	-4.3%	18.0%	1.81	199	80%
16.56	16.5	273	108	-6.1%	19.0%	2.02	218	80%
16.55	18.5	306	105	-8.7%	19.5%	2.33	245	80%

Table 2. Results of tests carried out on the breadboard implementation. Column headings: DC-V is the DC voltage supply into the inverter; DC-A is the DC supply current; Pin-W is the DC power input; AC-V is the RMS output voltage; ∆V is the percentage change in RMS output voltage when the load is varied according to the RMS output current AC-A, power dissipated in the load Pout W and operating efficiency. All AC measurements were made using true RMS meters for voltage and current [10] to account for the non-sinusoidal nature of the waveform.