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DDS RF SIGNAL GENERATOR

50 Hz to 71 MHz



Induction-
Balance Metal
Detector



Valve
Headphone
Amplifier

SuperCap
Battery

Xilinx PROM
Programmer

TinyLogic ICs

The P89LPC900



DDS RF Signal Generator

Frequency range: 50 Hz to over 70 MHz

Design by G. Baars, PE1GIC.

pe1gic@amsat.org

Radio amateurs and RF engineers in general will welcome this design. This contemporary RF signal generator has many bells and whistles and is just the ticket for testing HF/VHF receivers, aligning filters, IF amplifiers and AM/FM demodulators. The instrument can even act as a source for very low frequencies starting at 50 Hz.



While the combination of an ordinary multimeter and a rudimentary signal tracer is perfectly adequate for many 'kitchen table' hobbyists to get their homebrew circuits working, more sophisticated equipment is typically required by those interested in radio and higher frequency circuits in general. In particular, the more complex designs in the RF realms normally require accurate adjustment, which in practice translates into a decent RF signal generator with an internal or external modulation option available. But then, most of you will agree, we are talking of an expensive piece of test equipment and that is why we expect the present design to be highly valued, not only by the radio amateur fraternity but also by those with an interest in all things RF. The generator described in this article offers good performance, may be used for commonly required test, service and repair jobs, and is reasonably simple to build.

Concept and block diagram

The general design of the RF Signal Generator is shown in **Figure 1**. At first blush you might think that this is another circuit with a microcontroller at the heart of things, but this time the term is more appropriate for the module marked 'DDS', because this is where the RF signals are actually generated.

DDS is an abbreviation of 'Direct Digital Synthesizer'. The DDS requires a clock signal for its frequency reference. This signal is frequency-multiplied by six by the DDS. In this way, by applying a clock frequency of 30 MHz to the DDS, the internal clock frequency becomes 180 MHz, which is also the highest frequency at which the DDS can operate. Its sinewave-shaped output signal has a frequency f_0 equal to

$$f_0 = W \times (f_{\text{clk}} / 2^{32})$$

where W is a 32-bit programmable 'frequency word'. Consequently the step size becomes

$$180 \text{ MHz} / 2^{32} = 0.0419 \text{ Hz}$$

By means of software, the smallest step is set to a more familiar value, namely 1 Hz.

Because of the internal design of the DDS, a number of spurious signals are inevitably generated, particularly since the output signal is quantified at 180 MHz, there's no way to avoid a filter (module 'LPF' in the drawing). The filter applied here is a Butterworth low-pass type that's guaranteed to afford sufficient suppression of unwanted products.

An adjustable attenuator is required if we want to be able to control the output signal level. That is why the filter is followed by a digitally controlled VGA (variable gain amplifier). Using this VGA the gain can be set in set in 1-dB steps over a range of 31 dB. The VGA in turn is followed by two attenuators of 32 dB and 64 dB respectively. The total attenuator arrangement allows the output signal to be adjusted between 0 dBm (decibel milliwatt) and -127 dBm. The VGA used here doubles as a 50- Ω output signal driver.

Specifications

- Output frequency adjustable between 50 Hz and 71 MHz
- Frequency step size 1 Hz to 1 MHz
- Output signal level adjustable between 0 and -127 dBm (0.224 V_{rms} to 0.1 μ V_{rms})
- Internal AM, 1000 Hz at 30%
- Internal FM, 1000 Hz, deviation 3 kHz, 10 kHz or 20-90 kHz
- 16-key keypad for frequency entry and other functions
- 2x16 character LCD showing frequency, frequency step and output signal level
- Spurious output level -40 dBc to -50 dBc (frequency dependent)
- Frequency range covers standard IFs like 445 kHz, 5.5 MHz, 10.7 MHz, 21.4 MHz, 45 MHz and 70 MHz

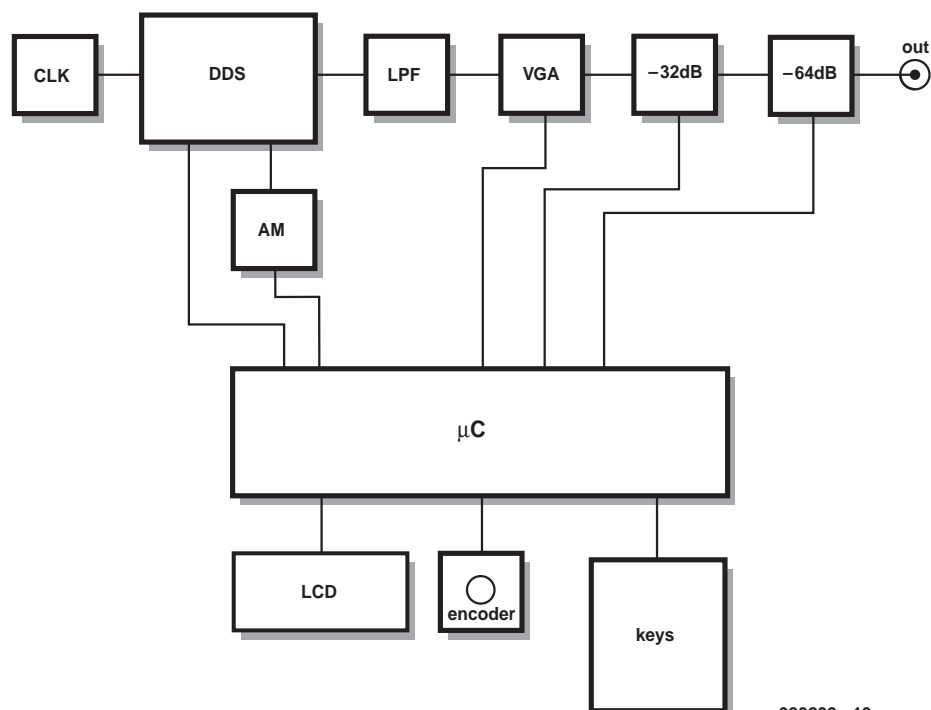
Arguably a microcontroller is the best choice, if not indispensable, if we want to control all of the above circuitry. Here, the micro rules over the DDS, attenuators and the 'user interface', the latter consisting of a keypad and LCD and a rotary encoder.

Circuit diagram

As you'll discover a bit further on, the circuit is spread across two printed circuit boards. The division is reflected by the circuit diagram. Broadly speaking, the schematic in

Figure 2 shows the signal generator proper, while the control circuitry, power supply and user interface appear in **Figure 3**.

The various elements discussed in relation to the block diagram are easily found back in the actual schematics. In Figure 2, the clock generator is built around IC1, while the DDS lurks in IC2. The low-pass filter is found around L6-L15, the VGA is integrated into IC3 and the attenuators are situated around relays Re1, Re2 and Re3. Connectors K1 and K2 are interconnected with Figure 3's K2 and K3 respectively. In Figure 3, most of you will immediately spot IC2 as the microcontroller. The keypad is connected to K1, the LC display to K4. S1 is a rotary encoder while Tr1,



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Figure 1. Block diagram of the DDS RF Signal Generator.

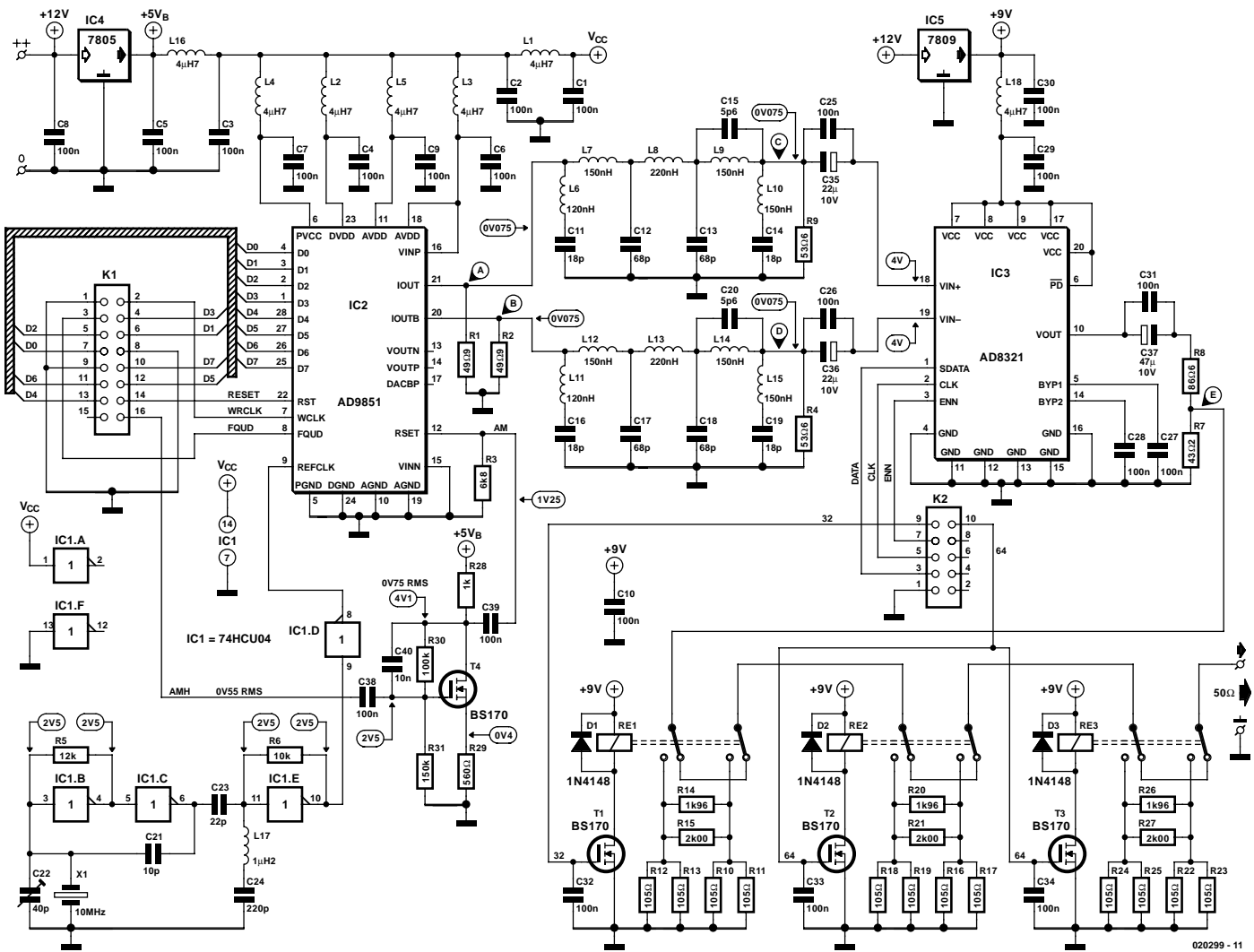


Figure 2. Circuit diagram of the RF part of the generator.

B1 and IC3 are the main parts in the power supply. The circuit around IC1 is part of an AM modulator and only happens to sit here because it could not be accommodated on the other board (Figure 2).

The operation and functionality of each of the above circuit sections will be discussed in the following paragraphs.

Clock generator

Because a 30-MHz TCXO can only be obtained as an expensive custom-made component, and 30.000 MHz quartz crystals are few and far between, a solution was found in the use of a 10-MHz oscillator in combination with a tripler. In this way we're able to employ a cheap and commonly available quartz crystal while the oscillator frequency is easily adjusted (here, with C22). Of course, the oscillator is not totally immune to temperature variations, but in practice it will operate satisfactorily because the equipment

will typically be used at room temperature.

IC1b, IC1c and surrounding components form the oscillator. The frequency tripler is built around IC1e, while IC1d acts as a buffer.

The combination of X1 and IC1 may be replaced with a 14-way DIP 30-MHz quartz oscillator module (if you can get it) which may be plugged into the socket for IC1. Unfortunately, 8-pin oscillator blocks will not fit the board and require a small modification which you will have to work out for yourself.

DDS

The circuitry around IC2 largely follows the application suggestions supplied by the manufacturer. The DDS chip has several voltage connections, each supplying its own

part of the complex chip. To keep spurious signal levels to a minimum, all supply connections are powered via a separate supply filter consisting of a choke and a decoupling capacitor.

The resistor at pin 12 of the DDS (here, R3) serves to defines the DDS output current. However, by allowing an audio signal to vary this resistance, AM modulation is obtained.

Filter

Inherent to its design, the DDS generates not only the desired frequency f_0 , but also the spurious products f_{clk} , $f_{clk}-f_0$ and multiples of these. Arguably a good filter is in order to keep spurious levels at the output as low as possible. The filter used here is a modified Butterworth low-pass dimensioned for a roll-off frequency

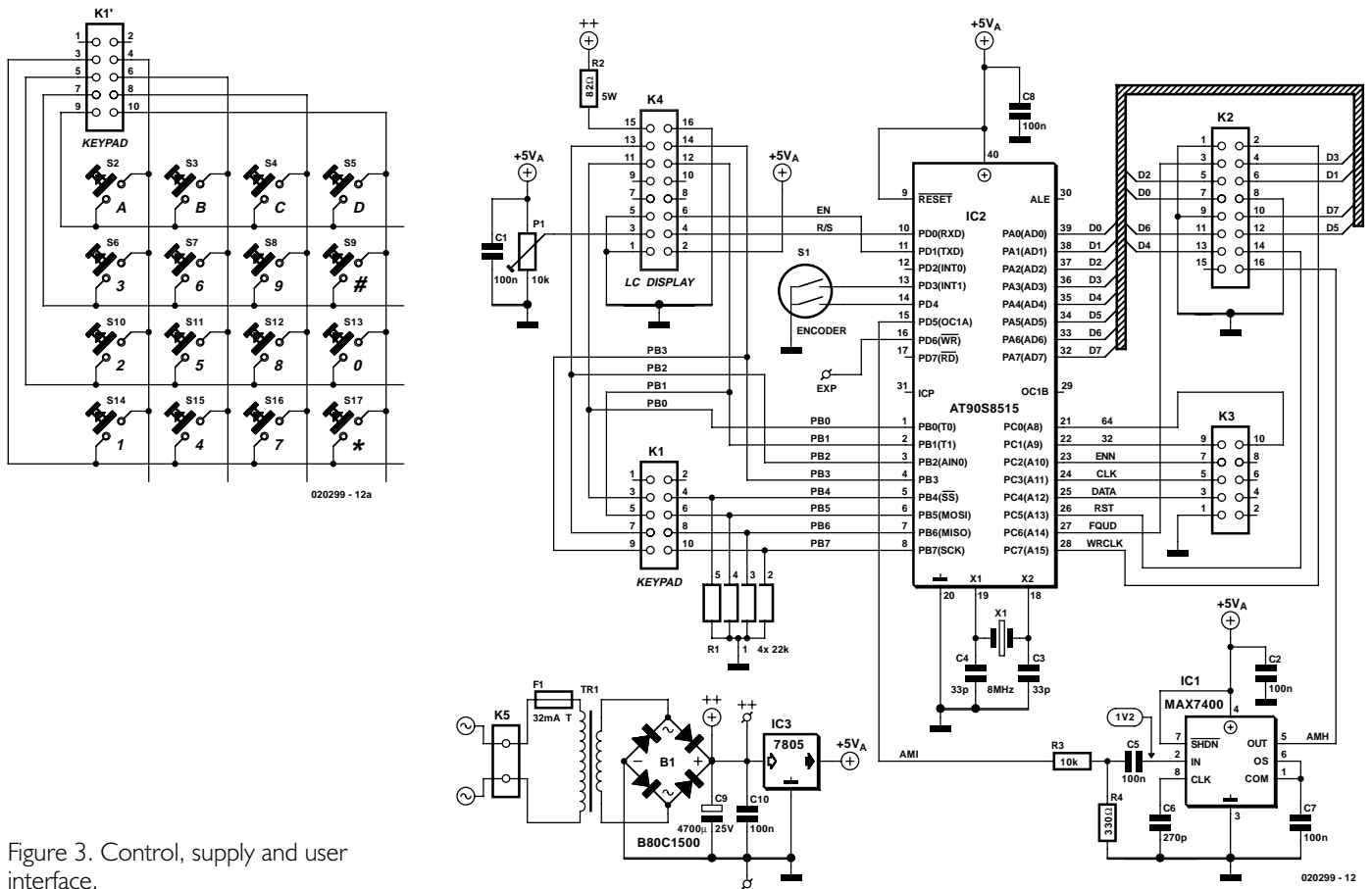


Figure 3. Control, supply and user interface.

of about 75 MHz. Since at an output frequency f_0 of 71 MHz the component $f_{\text{clk}} - f_0$ occurs at 109 MHz, the filter needs to have a fairly steep slope. This condition is fully satisfied by the filter used here because it is almost flat between 0 and 71 MHz, while a minimum suppression of 50 dB is obtained for signals above 95 MHz or so.

The filter has a double implementation because both outputs of the DDS are required to obtain symmetrical AM modulation.

VGA

IC3 comprises a digitally controlled variable-gain amplifier/attenuator with a flat amplitude characteristic over the entire frequency area we're concerned with. The gain is adjustable in steps of about 0.75 dB across a range a bit larger than the required 31 dB.

Because the VGA is intended as a 75- Ω driver and strives to dynamically maintain its output impedance, the relevant chip output must be terminated into 75 Ω . In our circuit, a couple of resistors are used to create

a level converter from 75 Ω to 50 Ω . None the less, the VGA is easily capable of generating 0 dBm which equals 1 milliwatt or 0.224 V_{rms} into 50 Ω .

Attenuators

Additional attenuators are called for if we want the generator to supply a lowest output level of -127 dBm into 50 Ω (which equals 0.1 μV_{rms}). The VGA itself already provides for 31 dB of attenuation, so we need to add a

Meet the DDS

The DDS consists of three parts. First, we have an NCO (Numerical Controlled Oscillator), which in the case of the AD9851BRS is a 32-bit counter that adds a 32-bit frequency word on every clock pulse. A small value for the 'word' causes the counter state to increase slowly — a larger value, quickly. By sending the 10 MSB's of this counter to a DAC via a function called Sine Look Up Table, a sinewave-shaped output voltage is created of which the frequency is variable.

Because the output voltage is quantized at the clock frequency, unwanted products are generated including the component $f_{\text{clk}} - f_0$. Obviously when $f_0 = 1/2 f_{\text{clk}}$ we have the case where $f_{\text{clk}} - f_0$ will actually equal f_0 . As a result, the highest usable output frequency of a DDS is usually limited to about 40% of the clock frequency. If not, a low-pass filter is required to give sufficient suppression of the unwanted product.

Another disadvantage of the use of a DDS is that its output level is not constant. In fact, the level is described by a $(\sin x/x)$ curve with $x = \pi \times f_0 / f_{\text{clk}}$.

Some more calculation indicates that $(\sin x/x)$ equals 0.76 or -2.4 dB at $f_0 = 0.4 f_{\text{clk}}$. While the error is not grave in a receiver where the DDS is used as a local oscillator, it is rather worrying in the case of an RF signal generator. Consequently, the VGA output level is corrected at various output frequencies. For this advanced function a special routine is implemented in the control software.

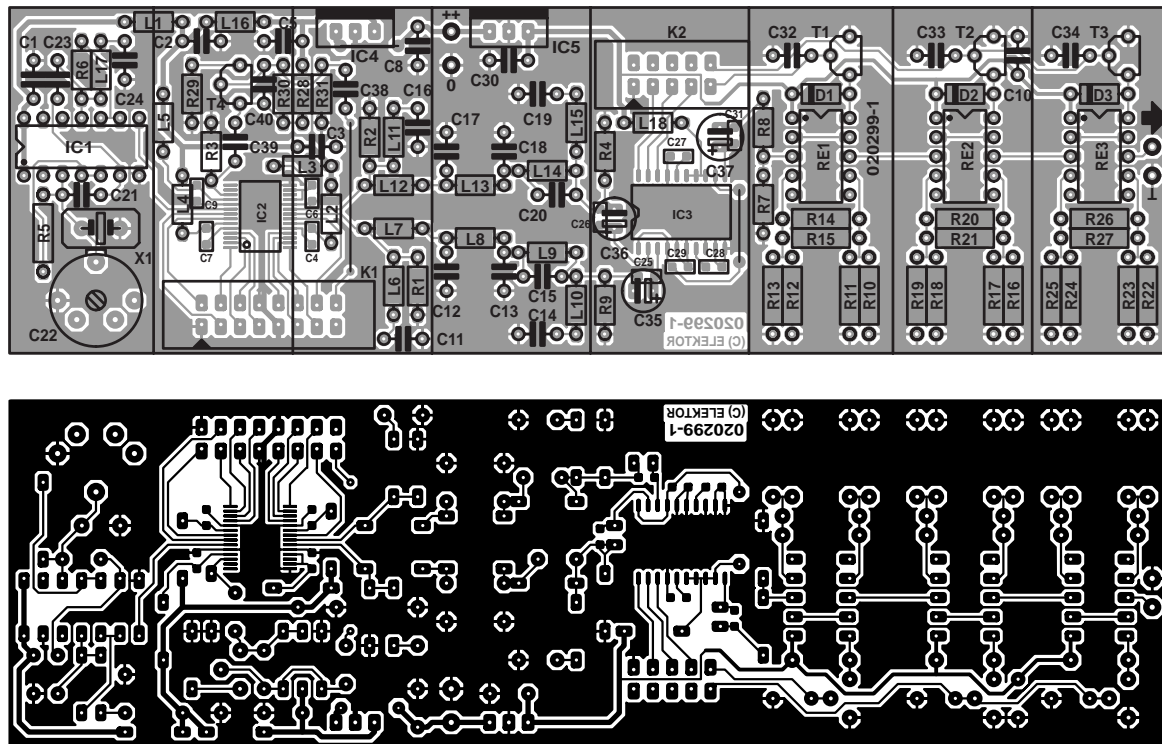


Figure 4. The PCB for the signal generator circuitry is marked by very short connections and a generous earth plane.

32-dB and a 64-dB attenuator. The latter comprises two series connected 32-dB sections, which are easier to produce in practice without the risk of inaccuracy or leakage associated with a single 64-dB attenuator.

The relays used here (Re1, Re2 and Re3) are configured to switch attenuator sections in and out of circuit under microprocessor control. Despite their relatively low price, the relays are usable for frequencies up to 1 GHz. In practice, these 12-volt models pull in reliably a just 9 volts coil voltage. In case of doubt, the relay supply voltage may be taken from the unstabilised +12-V rail in the circuit.

Control and AVR

The various circuits in the RF Signal Generator are controlled by an Atmel AT90D8515 microcontroller (IC2 in Figure 2). This 8-bit RISC controller offers 32 I/O lines and a speed of 8 MIPS which makes it perfect for the job. Parallel driving of the DDS guarantees that its programming is fast enough to modulate sufficient samples when FM is used. Note, however, that this does require 11 I/O lines. To save some I/O resources, the LCD and the keyboard share a number of processor pins.

The rotary encoder drives an interrupt line to make sure the software can not miss any pulse. The LCD is used in 4-bit bus mode where data is copied to it in two operations. That, too, is done to save I/O line capacity. Finally, sharing I/O lines between the DDS

COMPONENTS LIST

Signal generator board (020299-1)

Resistors:

- R1,R2 = 49Ω
- R3 = 6kΩ
- R4,R9 = 53Ω
- R5 = 12kΩ
- R6 = 10kΩ
- R7 = 43Ω
- R8 = 86Ω
- R10-R13,R16-R19,R22-R25 = 105Ω
- R14,R20,R26 = 1kΩ
- R15,R21,R27 = 2kΩ
- R28 = 1kΩ
- R29 = 560Ω
- R30 = 100kΩ
- R31 = 150kΩ

Capacitors:

- C1,C2,C3,C5,C8,C10,C30,C32,C33,C34,C38,C39 = 100nF, 5mm lead pitch
- C4,C6,C7,C9,C25-C29,C31 = 100nF, SMD shape 0805
- C11,C14,C16,C19 = 18pF
- C12,C13,C17,C18 = 68pF
- C15,C20 = 5pF
- C21 = 10pF
- C22 = 40pF trimmer
- C23 = 22pF
- C24 = 220pF
- C35,C36 = 22μF 10V radial

- C37 = 47μF 10V radial
- C40 = 10nF

Inductors:

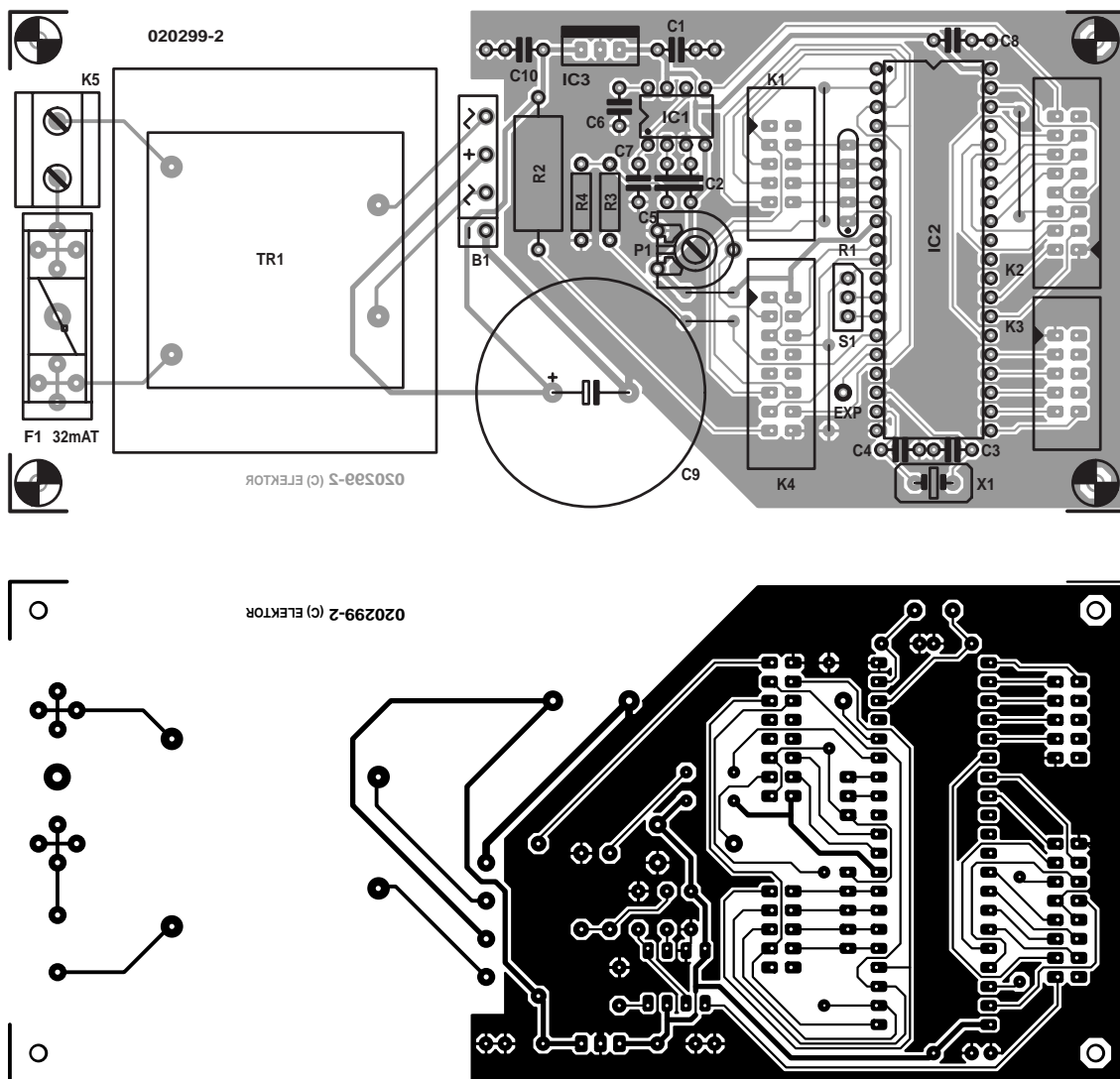
- L1-L5,L16,L18 = 4μH
- L6,L11 = 120nH
- L7,L9,L10,L12,L14,L15 = 150nH
- L8,L13 = 220nH
- L17 = 1μH

Semiconductors:

- D1,D2,D3 = 1N4148
- T1-T4 = BS170
- IC1 = 74HCU04
- IC2 = AD9851BRS
- IC3 = AD8321AR
- IC4 = 7805
- IC5 = 7809

Miscellaneous:

- K1 = 16-way boxheader (2x8)
- K2 = 10-way boxheader (2x5)
- X1 = 10MHz quartz crystal (series resonance. C_L 32pF) or 30MHz DIL14 oscillator module
- Rel,Re2,Re3 = TQ2-9V or TQ2-12V 3 wire links
- Enclosure, tin sheet dim. 160x48x25 mm
- PCB, order code **020299-1** (see Readers service page)



COMPONENTS LIST

Control/power supply board (020299-2)

Resistors:

R1 = 4-way 22k Ω SIL array
 R2 = 82 Ω 5W
 R3 = 10k Ω
 R4 = 330 Ω
 P1 = 10k Ω preset

Capacitors:

C1, C2, C5, C7, C8, C10 = 100nF, 5mm lead pitch
 C3, C4 = 33pF
 C6 = 270pF
 C9 = 4700 μ F 25V radial

Semiconductors:

B1 = B80C1500, rectangular case (80V piv, 1.5 A)
 IC1 = MAX7400CPA

IC2 = AT90S8515 8PC, programmed, order code **020299-41** (see Readers Services page)
 IC3 = 7805

Miscellaneous:

K1, K3 = 10-way boxheader (2x5)
 K2, K4 = 16-way boxheader (2x8)
 K5 = 2-way PCB terminal block, lead pitch 7.5mm
 S1 = rotary encoder, Bourns ECW1J or ddm427 (Conrad Electronics)
 X1 = 8MHz quartz crystal, parallel resonance, C_L 32pF
 TR1 = 12V/4.8VA mains transformer, e.g., Gerth 1x12V/400 mA
 F1 = fuse, 32 mA with PCB mount fuse holder
 5 wire links
 Keypad: 16 keys, matrixed (Velleman)
 Display: LCD 2x16 characters with backlight
 PCB, order code **020299-2** (see Readers Services page)

Figure 5. The supply/control board is much more spacious.

and the VGA is really out of the question with the risk of increased spurious levels in mind.

Keyboard, display and encoder

The user interface designed into the RF Signal Generator consists of a 2 x 16 character matrix LCD, a 4 x 4 matrix keyboard (connected to K1 in Figure 3) and a rotary encoder (S1). The LCD connected to K4 provides a readout for frequency, frequency step and output level. The keypad allows the desired frequency to be entered as well as various other functions to be controlled. The rotary encoder is used to adjust the signal frequency, select the frequency step size and adjust the output signal level.

The LCD backlight current is limited to a

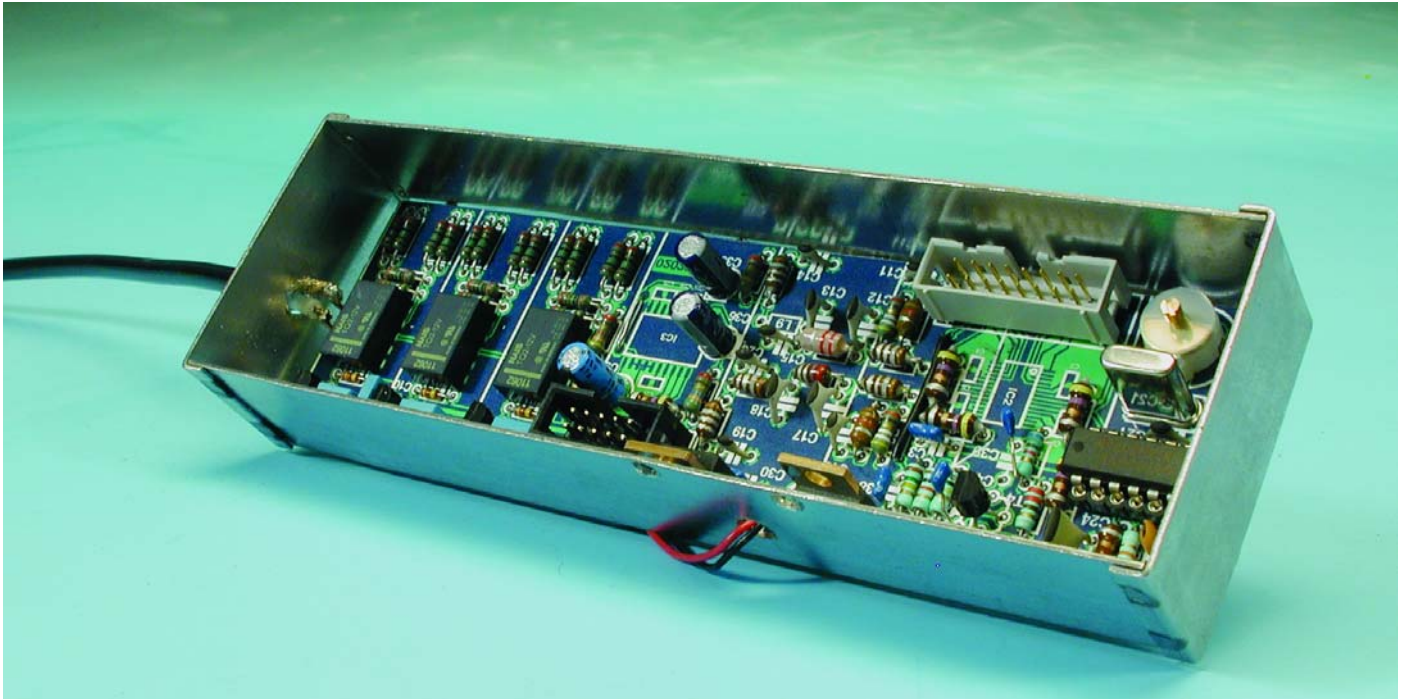


Figure 6. The signal generator board is by no means easy to construct.

safe value using series resistor R2. In practice, one third of the recommended current guarantees a sufficiently bright display. The saving in current then amounts to 200 mA!

Preset P1 acts as the LCD contrast adjustment.

AM modulator

Most RF signal generators of the affordable kind use fixed 30%, 1000-Hz AM modulation. Because the DDS has no internal provision for amplitude modulation, an external add-on had to be devised. The resistance at pin 12 of the DDS determines the DDS output level. By using a FET (T4), this resistance can be varied dynamically. The sinewave applied to the FET is obtained by filtering a square wave from processor pin 15. The filter in question is a pretty steep one, built around a Maxim integrated elliptic low-pass (IC1 in Figure 3). The filter suppresses the fundamental frequency of the square wave, which results in a clean 1-KHz sinewave.

FM modulation

Frequency modulation (FM) is realised in software, with the microcontroller employing an internal processor timer and a sinewave look-up table containing frequency steps. FM with 1000-Hz sinewave modulation is obtained by sending 32 samples to the DDS at a timer rate of 32 kHz. The number of samples and the sampling frequency distance are large enough to warrant a reasonably clean modulated spectrum.

The keyboard allows you to select FM modulation with a deviation of 3 kHz, 10 kHz, 20 kHz, 30 kHz and so on up to 90 kHz. The 3-kHz setting will typically be used for NBFM equipment like personal mobile radios, while 70 kHz is the nominal value for broadcast FM.

The sinewave tables for the FM modulation function of the instrument have been developed using a specially written Pascal program.

Software

The microcontroller executable code was created using an assembler program with well over 2,000 lines. Broadly speaking, this program consists of three flows:

Main flow

In the main flow, the keyboard is scanned and the keyboard presses are linked to their associated functions. From the main flow, a number of subroutines are called controlling, among others, the LCD. Here, too, the interrupt timer is initialised for AM and FM.

The main flow is preceded by a reset interrupt which arranges for all hardware and software initialisations to be carried out.

Timer interrupt

The timer interrupt is activated at a rate of 2,000 Hz or 32,000 Hz for AM and FM respectively. With AM, the timer interrupt causes a square wave to appear on an I/O pin. With FM, a frequency sample from the sinewave look-up table is added to the current frequency and the result is sent to the DDS.

Encoder interrupt

When an encoder interrupt occurs, either the frequency, frequency step size or the output level is increased or decreased. Next, besides other 'chores', the display readout is updated. The function of the encoder is determined using the keyboard.

Keyboard functions

The complete functionality of the instrument is accessible to the user via the keypad and the rotary encoder. The keyboard functions have been defined as follows:

- * rotary encoder controls frequency step size
- 0 rotary encoder controls output signal frequency
- # rotary encoder controls output signal level
- 0-9 output signal frequency

- D** enter output signal frequency
- A** modulation AM/FM/off
- B** attenuator display format dBm or V
- C** FM deviation: C0-C9 (C0 = 3 KHz; C1-C9 = 10-90 kHz)
- D** output signal on/off

Notes:

- The desired output frequency does not appear at the output until 'D' is pressed.
- If FM is selected using key 'A', the display will indicate 'F1'. After pressing 'C' (display: 'F?') the desired deviation may be entered using the number keys.
- By pressing '*' you can set the step size applied to the current frequency on the display, i.e., the increment/decrement caused by one click of the rotary encoder. The step size appears in the left-hand bottom corner of the display. The output level appears at the other side.

Power supply

The complete circuit draws up to 400 mA at a supply voltage of 12 V,

which allows a relatively small on-board mains transformer to be used. After rectification and smoothing, regulators are used to create the various supply voltage rails needed in the circuit. Each part of the circuit receives its own supply voltage: C3 on the control board looks after the microcontroller power supply, while IC4 and IC5 on the main board are the respective supplies for the DDS (plus clock oscillator) and the VGA.

Because the circuit has its own mains-connected power supply, due attention should be paid to electrical safety when assembling the electronics into a case. In particular, make sure a good strain relief is used on the mains cord. If desired the mains transformer may be omitted from the board in Figure 5 and replaced by a mains adaptor (battery eliminator) rated at 12 V / 0.5 A. Finally, a tip: a few turns of the mains cord through a ferrite cord will reduce RF leakage through the mains.

Construction

As already mentioned, the complete

circuit is spread across two printed circuit boards — one for the signal generator proper and another for the control and power supply sections. The first board corresponds to the schematic in Figure 2 and its artwork is shown in **Figure 4**. For the second board the correspondence is between the schematic in Figure 3 and the PCB artwork in **Figure 5**.

The combined supply/control board of Figure 5 has a spacious layout and contains conventional components only, so should be easy to build by anyone with some practical skills in DIY electronics. Do not forget to fit any of the five wire links on this board.

The printed circuit board pictured in Figure 4 is a different kettle of fish. With stability in mind and in order to keep stray radiation to a minimum the design of the board follows the 'great RF tradition' of short connections, the smallest possible lead pitch for components and a maximum amount of electrical separation between various part of the circuit.

Building the signal generator board requires care, precision, good soldering skills and a steady hand. After all, IC2 and IC3 are SMD devices, which also applies to a dozen or so coupling and decoupling capacitors around these integrated circuits. All these SMD parts are fitted at the **underside of the board**.

Soldering SMD components requires special skills. While IC3 is still relatively easy to handle, soldering IC2 in place could pose unexpected problems as the part has a pin spacing of just 0.65 mm. First carefully pre-tin the footprint of the IC and then remove as much tin as you can using fine desoldering braid. Use a drop of hobby glue to secure the IC in place. Use a magnifying glass to check that all the pins are properly aligned to the copper pads. If necessary, adjust the position of the IC and then allow the glue to cure.

With the IC firmly in place, first solder the centre two pins using plenty of solder tin and not caring too much about excess solder causing short-circuits. Allow the IC to cool down between subsequent solder actions. Once all pins are covered in plenty of solder tin, the excess amount can be removed by means of desoldering braid. Here, too, the IC should not be endangered by overheating so take your time.

The next step is to use an ohmmeter to check for short-circuits between adjacent pins. If any are found, re-apply the desoldering braid until no more short-circuits are found.

The photograph in **Figure 6** shows a finished prototype of the signal generator board while **Figure 7** zooms in on the underside of the board, showing the vicinity of IC2.

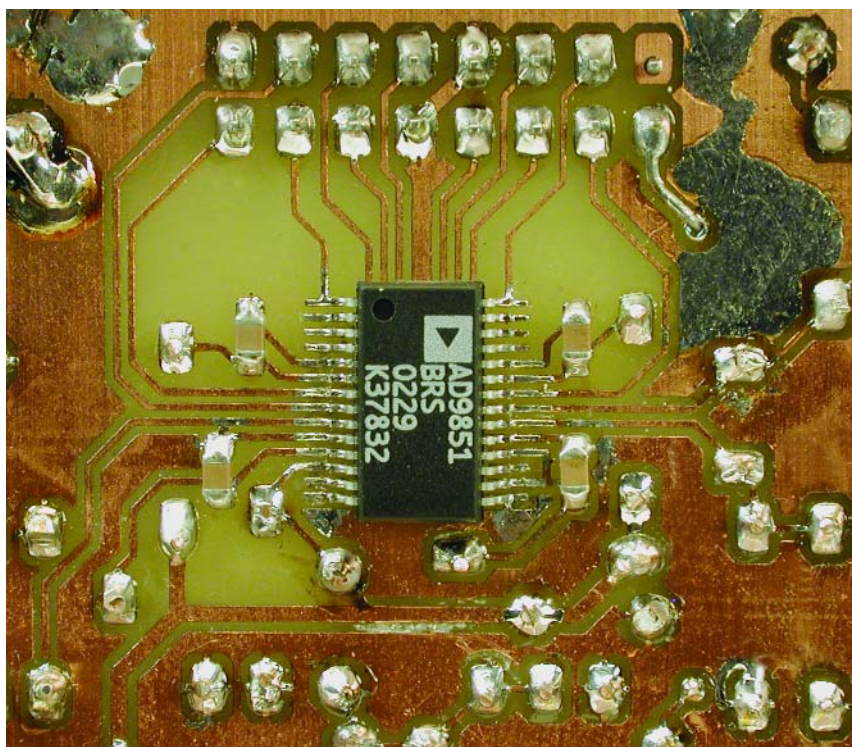


Figure 7. SMD integrated circuits IC2 and IC3 are fitted at the underside of the board, together with a dozen or so SMD passives. The mounting of IC2 is tricky owing to the small lead pitch of just 0.65 mm.

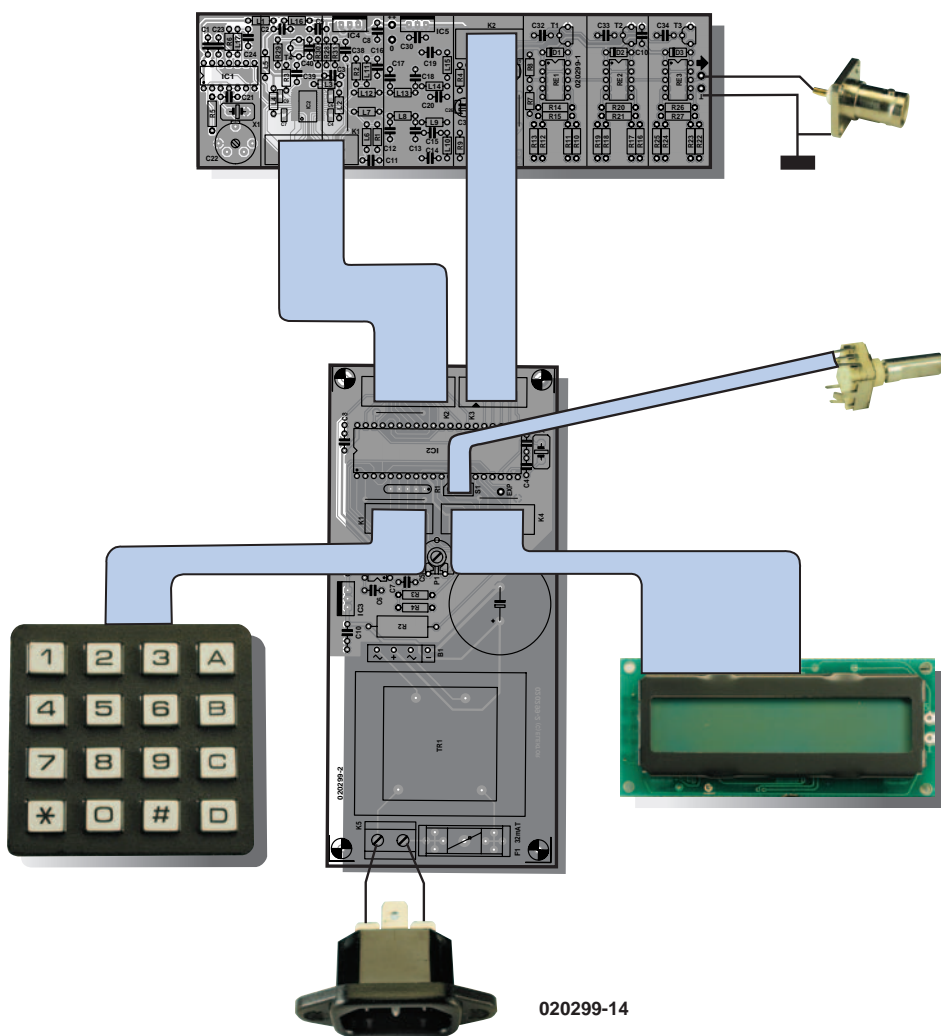


Figure 8. Wiring diagram.

Enclosure

For obvious reasons, a metal enclosure is a must for an RF signal generator and the project discussed here is no exception. The good news is that only a small enclosure is required, and suitable tin-plated steel cases with detachable lids are available in several sizes. In our case (pun intended), the required size is 160 × 48 × 25 mm. The signal generator board allows small metal screens to be fitted to separate the various circuit sections. The position of these screens is indicated by lines on the component overlay. In the case of our prototype, no differences were measured with the screens in place or removed, so we decided to leave them out. Perfectionists are, of course, free to fit whatever screens they think are necessary.

Next, the various units that make up the equipment may be assembled together allowing the complete signal generator circuitry to

be mounted in a suitable case. Our prototype boards were fitted in a 'custom' case made from pieces of unetched circuit board of which the copper surfaces were connected by soldering. Electrically, such a DIY case is equivalent to one made from metal sheet.

As a further aid to your own construction work, **Figure 8** provides a basic wiring diagram showing how the two boards, the LCD and the rotary encoder are interconnected.

In the unlikely case of text failing to appear on the display immediately after switching on, do not panic and start emailing us, but first adjust the LC contrast control, P1. If the rotary encoder appears to 'turn the wrong way around', simply swap the wires to the two outer connections — the centre connection is ground.

Components

The author obtained all components for this project from RF specialist Barend Hendriksen in Brummen, The Netherlands (www.xs4all.nl/~barendh/Indexeng.htm).

The keyboard used in the prototype was a 16-key matrixed type supplied to us by Velleman (www.velleman.be). The LCD is an industry-standard 2 × 16 character alphanumeric type (Mitsubishi, Hyundai, etc.). The rotary encoder, finally, is a Bourns type with 24 detents per revolution.

Sensitivity measurements

In many cases, an RF signal generator will be used to measure the sensitivity of a receiver or IF amplifier. Usually, you'll want to know the sensitivity in microvolts (μV) at a certain signal-to-noise ratio. The following method may be applied to obtain meaningful measurement results with a minimum of effort.

Connect the RF Signal generator to the receiver input by means of a short length of good quality 50-Ω coax cable like RG58C/U and ditto plugs. Switch off any computer equipment which is prone to leak spurious radiation into the receiver. Adjust the signal generator to the receiver frequency and then switch off the test signal by pressing the "D" key. Next, with the receiver volume control sufficiently 'up', use a multimeter or an oscilloscope to measure the level of the AF noise produced by the receiver. Switch on the test signal again and increase the output level from the lowest point (-127 dBm) to a level at which the AF noise level has dropped to a quarter of the that without an input signal. The difference represents a signal-to-noise ratio of $20 \log_{10}(4) = 12 \text{ dB}$, or "12 dB SINAD". To obtain the 20-dB SINAD sensitivity value for your receiver, increase the generator output level until the noise voltage has dropped to 1/10th.

(020299-1)

Valve Preamplifier

Part 2: construction

Design by Bob Stuurman

In the first part of this article, we described all of the circuit boards. Now it's time to look at how they fit together. We first describe the mechanical construction, after which we turn our attention to wiring the entire assembly. We conclude with a brief look at the specifications and performance.



An important factor with this type of equipment is the enclosure. For the prototype, we selected a standard model from the Conrad line. It consists of black top and bottom covers made from robust 1.5-mm sheet steel and two 1-mm aluminium face panels. The top and bottom covers have ventilation openings. The edges of the face panels are bent at right angles to form lips, to which the top and bottom covers are fastened using four self-tapping screws. The face panels have a brushed matte finish and protective cover foils. For the sake of simplicity, we call the face panel at

the front the 'front panel' and the other face panel the 'rear panel'.

Mechanical construction

The construction chosen by the author is the peak of simplicity. It uses a U-shaped channel section formed from 1.5-mm sheet aluminium, with dimensions of 290 × 155 × 85 mm (**Figure 1**). All of the preamplifier subassemblies are fitted on or inside this channel sec-

tion. Four 15-mm brass standoffs are fitted to the front, and another four 20-mm standoffs are fitted to the rear. The face panels of the enclosure are attached to the standoffs using M3×5 screws. The end result is that the aluminium channel section is 'suspended' between the front and rear panels.

There is a 15-mm space between the front face of the channel section and the front panel. This provides room for the fixing nuts of the potentiometers and switch and the gear-wheels of the balance potentiometers. The LED indicator lamp and the actuating button for the mains switch are fitted directly to the front panel. At the rear, the space is 20 mm deep. This region holds the I/O board (on 15-mm standoffs), the mains filter and IEC appliance receptacle and a small fan.

The power supply is fitted against the back of the channel section, with the transformer stack in the middle, the low-voltage circuit board to the left and the high-voltage circuit board to the right. The two amplifier boards are fitted at the front of the channel section. All of these circuit boards are screwed to 10-mm standoffs, with plastic standoffs being used for the high-voltage board. An aluminium plate with a thickness of 1 mm and a height of 55 mm, with a turned-up lip along one edge, is placed between the power supply

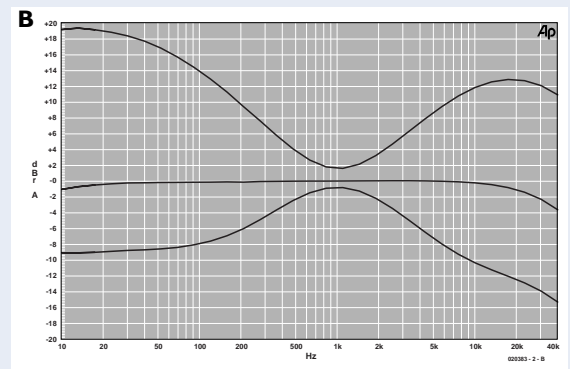
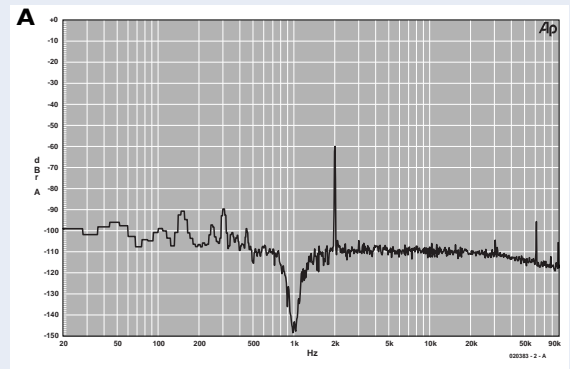
Specifications

Nominal signal level, 'Out'	450 mV (380 mV in)
Nominal signal level, 'Line Out'	220 mV (380 mV in)
Input impedance (CD/Tuner/...)	120 k Ω
THD+N, 'Out'	0.1 % (450 mV out)
Signal-to-noise ratio, 'Out'	80 dBA
Bandwidth (maximum volume)	< 10 Hz–35 kHz (–3 dB)
Crosstalk	< –65 dB (1 kHz) < –40 dB (20 kHz)
Channel separation	> 76 dB (1 kHz) > 60 dB (20 kHz)
Phono sensitivity (450 mV out)	5 mV (1 kHz)
Signal-to-noise ratio, Phono	> 53 dB
Bass adjustment (theoretical)	+18/–9 dB (50 Hz)
Treble adjustment (theoretical)	+9/–10 dB (10 kHz)

Measured response curves:

Chart A shows the frequency spectrum at maximum volume. The distortion primarily consists of the 2nd harmonic at –60 dB, which explains the value of 0.1 % for THD+N. The fundamental was suppressed for the measurement. The supply-voltage ripple and induced noise from the transformers lie below –90 dB and are negligible.

Chart B shows the frequency response with the tone controls in the neutral, minimum and maximum positions. The actual frequency response may vary due to component tolerances (potentiometers and capacitors).



and the amplifier boards. The lip of this plate is fitted underneath the standoffs for the amplifier boards in order to securely attach it to the channel section. In the prototype, a length of plastic cable duct (the type with a hinged snap cover) was fitted along the top edge of this aluminium plate on the side towards amplifier boards, in order to provide a cable-way for several cables. This cable duct can be clearly seen on the photo in **Figure 2**, which also shows the placement of the circuit boards and other components.

Details

Cooling fan

The ventilation openings in the top and bottom covers are not sufficient for the amount of heat to be dissipated. We did not want to deface the enclosure by making large holes in it, so we fitted a small fan at the rear of the enclosure. A series resistor (82 Ω in the prototype) causes the fan to turn quite slowly, and since it is fitted using rubber bushings, it is practically inaudible. The air stream from

the fan is directed toward the heat sink of the LM317. This forced airflow through the enclosure also removes the heat generated by the valves.

After the front and rear panels have been firmly screwed to the standoffs, the fastening holes for the top and bottom covers will quite likely not align exactly with the corresponding holes in the face panels. This is because it is nearly impossi-

ble to form the U-shaped channel section to a precise dimension. Consequently, the standoffs at the rear must be lengthened using shim washers until the holes are aligned. This also prevents the fan from being clamped between the two surfaces, so it can do its job without making any noise.

On/off switch

In order to keep the mains wiring (a potential source of interference) as short as possible,

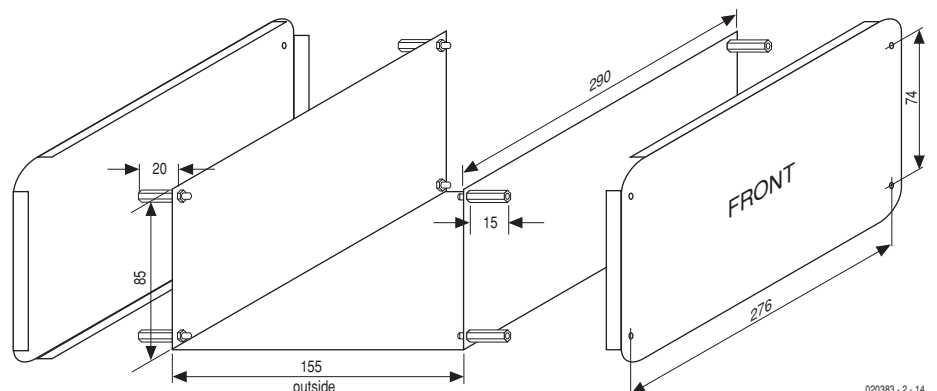


Figure 1. A U-shaped aluminium channel section is suspended between the front and rear panels of the enclosure, using 15-mm standoffs at the front and 20-mm standoffs at the rear.

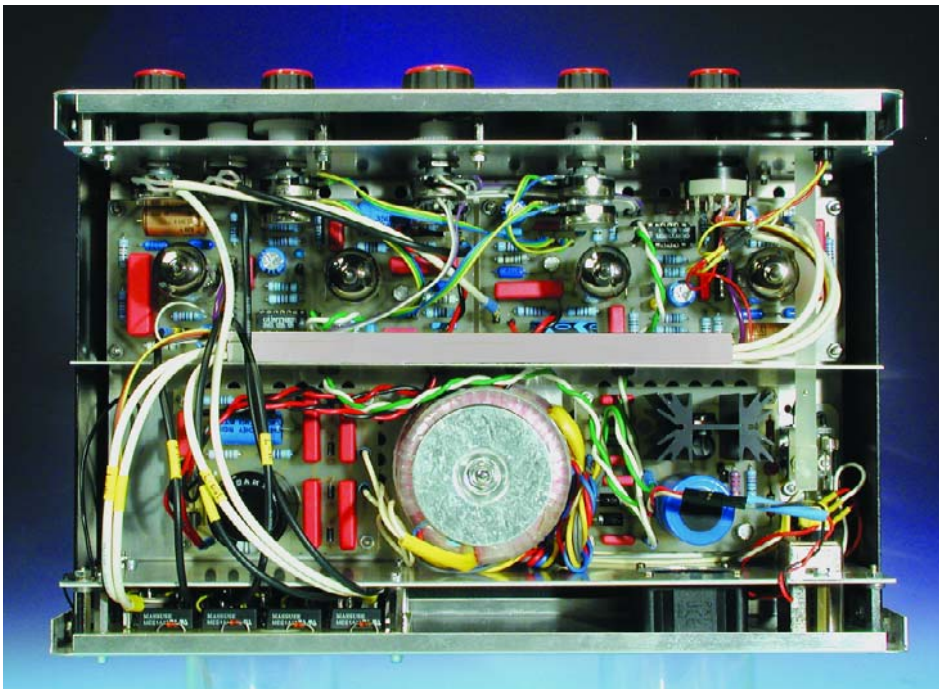


Figure 2. A 'bird's-eye view' of the inside of the preamplifier. All of the wiring is in place, the front and rear panels have been fastened to the channel section and the knobs are fitted.

the on/off switch should preferably be fitted as close as possible to the appliance receptacle. A pushbutton switch with a suitable mechanical extension reaching to the front panel is thus recommended. In the prototype, a special construction was used for this purpose, consisting of a rocker switch with a homemade extension arm, but we won't bore you with further details.

Balance control

As already mentioned in the first part of this article, the balance control is constructed using two log-taper potentiometers that are mechanically coupled using gearwheels (see

Figure 3). Drill and tap the gearwheels for M3 setscrews. As the material is quite soft, it's a good idea to file flats on the shafts of the potentiometers. This will allow the gearwheels to be securely anchored with only moderate tightening of the setscrews.

Next comes a tip: it's quite easy to equip the balance potentiometers with a tangible midrange position and 'click stops'. This can be done by fitting the shafts of the volume and tone control potentiometers with the same type of gearwheels as

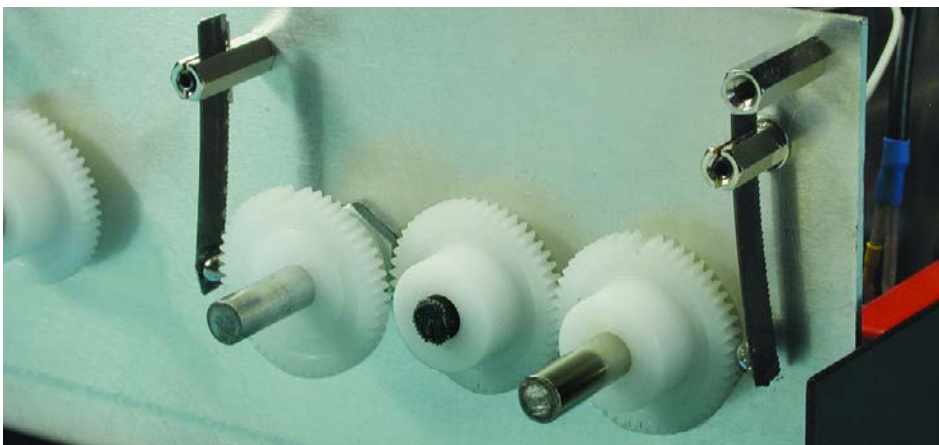


Figure 3. The balance potentiometers are coupled by a pair of gearwheels. The centre-to-centre spacing is 25 mm.

used for the balance control. If you arrange a leaf spring such that it presses a steel ball (from bicycle bearing, for example) against the teeth of the gearwheel, you obtain click stops. A tangible midrange position can be produced by removing one tooth at the midrange position. All of this is clearly shown on the photo in **Figure 3**.

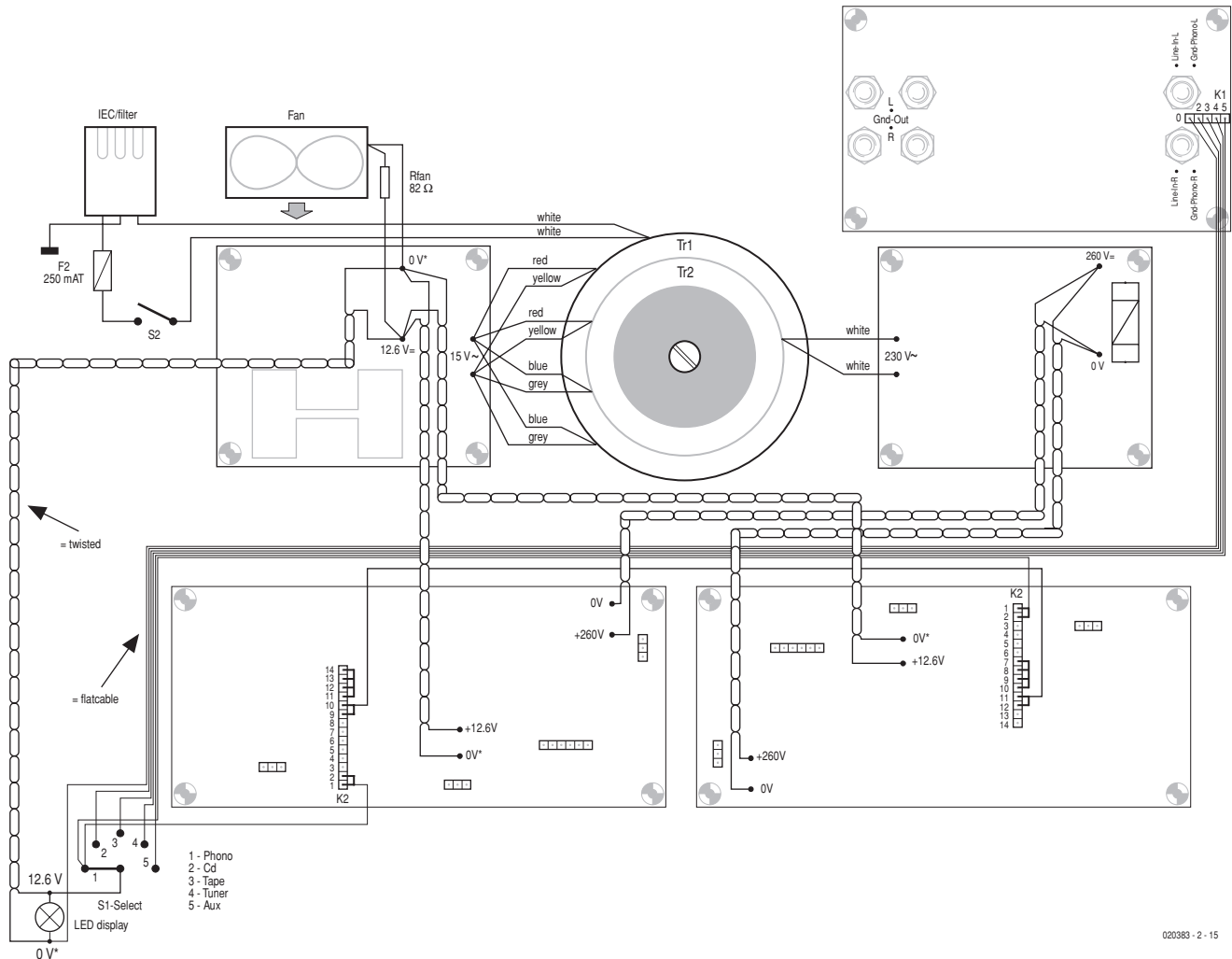
Front panel layout

You are naturally free to label the front panel with text and/or symbols according to your personal taste. For those who prefer a ready-made solution, a front panel layout is available on the *Elektor Electronics* website for download free of charge. It is also ideal for use as a drilling template for the front panel.

Wiring

The aluminium middle plate has four holes with feedthrough bushings, aligned with the circuit board connections for the filament voltage and high voltage, to allow the wiring to pass through the plate. These holes are located at half the height of the plate.

Figure 4 shows how the power wiring of the preamplifier should be fitted. Start by connecting the transformers to the power supply circuit boards, fitting the wiring for the cooling fan and making the mains voltage connections between the IEC appliance receptacle and the switch, fuse and mains filter. Connect the filter terminal to a solder lug screwed to the channel section. Next, make the connections to the filament and high-voltage terminals using twisted-pair wiring passing through the feedthrough bushings. Fit the 'filament' jumpers to the K2 connectors on the circuit boards and connect the filaments of the EF86s in series with a length of wire. This wire is routed through the cable duct. Finally, connect the LED indicator lamp and the input selector switch (S1). Terminal 1 of S1 connects directly to K2/1-2 on the left-hand amplifier board, and then to K2/1-2 on the right-hand amplifier board via a length of flat cable. This six-way flat cable (which is reduced to 5 leads before continuing to the I/O board) also runs through the cable duct.



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Figure 4. Connection diagram for the power wiring.

Figure 5 shows how the signal wiring must be fitted. The volume, treble and bass potentiometers are connected using short lengths of flat cable, while the rest of the connections are made using screened cable. It is convenient to first cut lengths of screened cable and connect them to the I/O board before it is screwed in place, since it is then still readily accessible.

Be particularly careful with the connections for the cable screens. At the Phono In connectors, the screen is connected to the ground point on the circuit board, and on the I/O board the same screen is connected to Gnd L or Gnd R, respectively, which is also the connection point for the screen of the Line In cable. For the output cables, the screens are jumpered to the circuit-board ground points at P5 on each amplifier board. At the balance poten-

tiometers, the screens are connected to the terminals on the left, and on the I/O board they are connected to Gnd Out, which is also the connection point for the screens of the Line Out cables. The three screened cables leading to the left-hand amplifier board run through the cable duct. The 'ground network' is connected to the aluminium channel section at one point only, which is Gnd L on the I/O board.

As can also be seen in Figure 5, the additional DC blocking capacitors Cx and Cy can best be soldered directly to the Line In sockets on the I/O board.

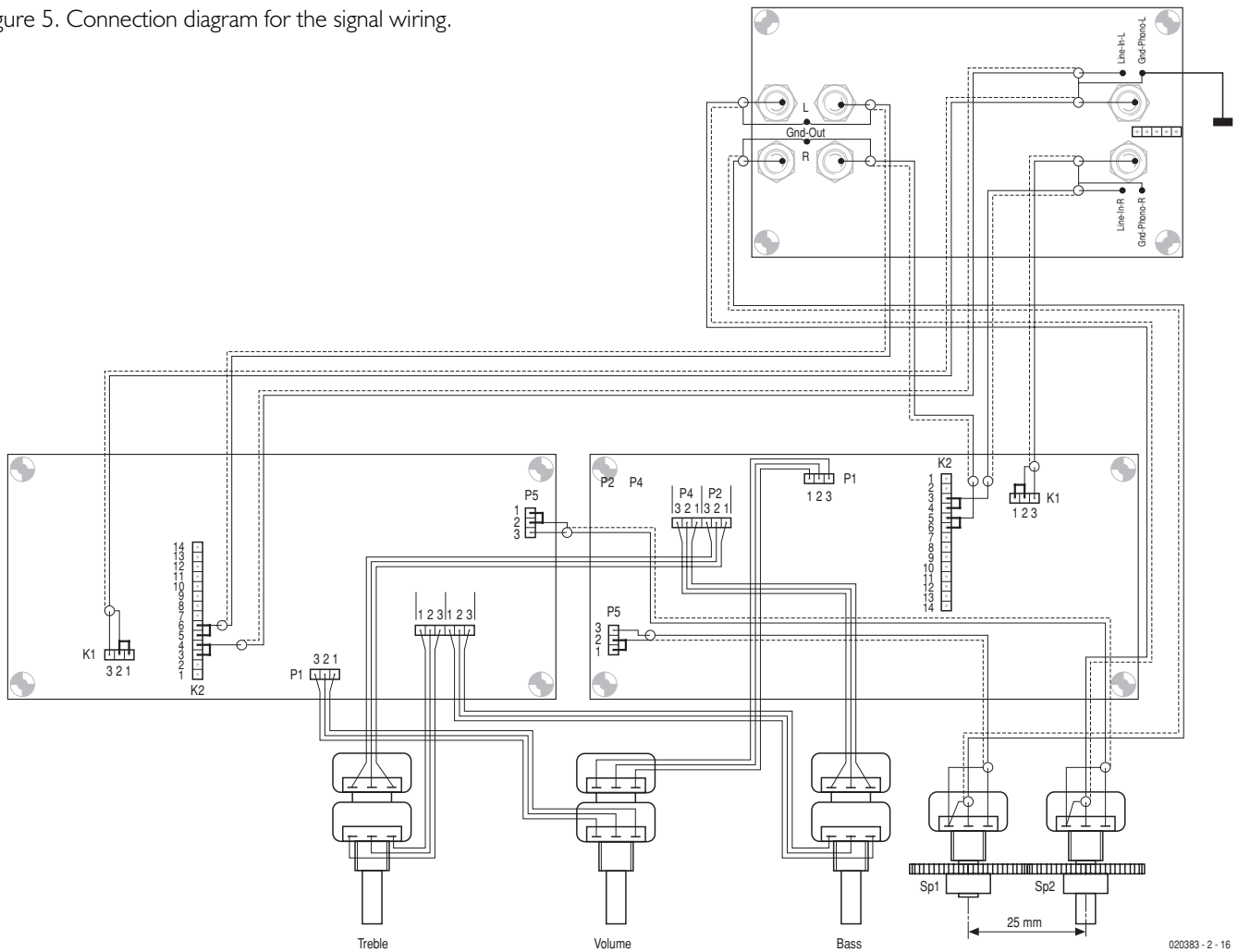
Alignment

Aligning the tone controls is best done by connecting an oscilloscope or AC millivoltmeter to the output. Turn P2 and P4 to the left and right

limits of travel in turn, and at each position, adjust P3 to make the minimum and maximum values for the left and right channels as nearly as possible the same (at 50 Hz and 10 kHz, respectively). Since log-taper potentiometers are used for the tone controls, the midrange position will not precisely coincide with a flat frequency response. In order to find the setting where the response is flat, apply a 1-kHz square wave signal to the input (not the Phono input) and observe the output signal on an oscilloscope. Turn the bass and treble controls until the square wave is as good as possible. With the potentiometers adjusted to achieve this condition, secure the knobs for the tone controls with their arrows pointing to '0'.

To align the balance control, start by setting the right-channel balance potentiometer (to which the knob will be attached) to its midrange position. Then rotate the other potentiometer (but not the gearwheel!) until the output signals of the two amplifiers have equal amplitudes. With the potentiometers in

Figure 5. Connection diagram for the signal wiring.



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this position, secure the gearwheel in place. If any small difference remains, it can be adjusted by slightly rotating the case of the potentiometer.

After the final (output) amplifier is connected, a slight amount of noise and some residual hum will be audible when the volume control is set to maximum and the Phono input is selected. When the volume control is rotated, you will probably hear a weak 'rasping' noise. This can be eliminated by using a small metal spring to electrically connect the shaft of the potentiometer to the enclosure. The hum will also disappear after the upper and lower metal covers of the enclosure have been fitted in place. The two halves of the enclosure can be connected to the central grounding point via lengths of flexible wire.

Connection to the final amplifier

Screened audio cable sometimes has rather high capacitance (as much as 200 pF per metre). If the distance between the pre-amplifier and the final amplifier is forced to be

rather large, it is worthwhile to give some attention to the connecting cables. In this case, you should select cables with low capacitance, so the high frequencies will be affected as little as possible. The author even built cables using RG-59

75-Ω coaxial cable (diameter 6.2 mm) and 'high-end' Cinch connectors. This type of cable has a capacitance of only 69 pF/m, so it can easily cover distances of up to several metres.

(020383-2)

MECHANICAL COMPONENTS LIST

- Two-section steel enclosure, 300x200x110 mm (wxdxh), Conrad Electronics # 520489*
- IEC appliance socket with internal filter
- Ventilator 40 x 40 mm, 20mm thick, 12V_{DC}
- 2 ABS gearwheels 50.M0.5, Conrad Electronics # 237850*
- S1 = rotary switch, 6 positions, 2 poles, break before make, Conrad Electronics 709751*
- S2 = mains switch
- F2 = fuse 250 mA (slow), with holder
- LED signal lamp, 12 V_{DC}
- 4 black buttons, 21mm
- 4 button caps, red, with line, 21mm
- 1 button, 28mm
- 1 button cap, red, 28mm

* www.int.conradcom.de

Xilinx PROM Programmer

Parallel port serial PROM burner

By Müller & Müller

The programmer described here reads and writes to the Xilinx family of serial PROMs. The PC software is written in Turbo Pascal and the programmer plugs into a free printer (Centronics) port.

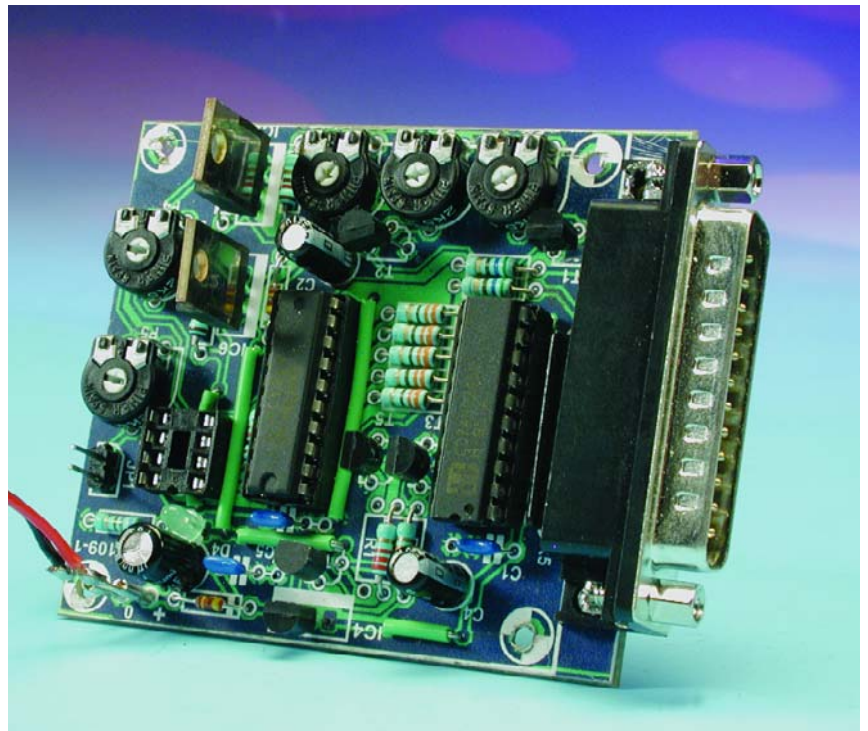
When it comes to PROM burning tools the most common model available is the universal type of PROM programmer. These are capable of handling the majority of chips currently on the market. The cost of such a burner is relatively high so the average electronics enthusiast has a hard job justifying his/her purchase especially if the programmer will only be used occasionally. If you do manage to scrape the funds to buy a programmer, regular software updates are fairly painless but it can be frustrating to find that a whole new model of programmer is introduced shortly after you make your purchase.

One solution to the problem is to build a low cost programmer and write (and update) the necessary programming software yourself. A disadvantage of choosing this path is that the programmer will not be so 'universal' as a commercial unit. The programmer described in this article is designed to burn serial PROMs produced by Xilinx but it should also be possible to adapt it for other manufacturers devices. These serial PROMs are used in conjunction with FPGAs (Field Programmable Gate Arrays). At boot-up the serial PROM loads configuration data to the FPGA. Together these devices are typically used to implement large complex digital systems (e.g., PCI bus designs).

The PROM burner described here plugs into a parallel printer port on a PC and the programmer software is written in Turbo Pascal and runs under DOS so even aged PCs should have no problem using it.

Burning Serial PROMs

The circuit diagram and PCB layout (Figure 3) are designed to accept PROMs with DIP out-



lines. For a more complete description of the devices refer to [1]. **Figure 1** shows the timing waveform specification for programming the XC17VXX-Family of PROMs.

As the name suggests, data to and from the PROM is transferred serially through a single pin (DATA) of the PROM. Writing begins at address '0000' with the data presented not in bits but in words. Most Xilinx PROMs have a 64-bit word length.

In order to switch the PROM into reading or writing mode it is neces-

sary to configure it with the correct bit pattern. These patterns are described under the heading *Programming* but for more detailed information see reference [2]. The timing diagram indicates the pattern and sequence of signals necessary for programming the PROM. Data is shifted into a 64 bit long latch in the PROM on the rising edge of the clock when both \overline{CE} and \overline{OE} are high. Once all 64 bits have been shifted into the register the voltage on pin V_{pp} is raised to V_{pp1} and the data is

latched. To read data from the PROM \overline{CE} is now pulled Low and \overline{OE} High. Data is shifted out of the PROM on every rising edge of the clock. Address incrementing occurs via the RESET/ \overline{OE} pin.

In addition to simple data transfers to the PROM there are also special features that allow the RESET polarity to be inverted and also some internal information to be read-out, for example the PROM type and manufacturers code (MfG code).

The Hardware

The programmer circuit diagram is shown in **Figure 2**. Printer port outputs IN00 to IN07 are used along with the two inputs ACK (Acknowledge) and PE (Paper Empty). A standard printer extension lead connects a PC to the 25-way plug K1 on the programmer. Outputs from the PC are used to switch the voltages in the programmer and also to provide the data, clock and reset signal to the PROM. The two inputs to the PC carry the PROM output data and the status signal CEO.

Signals on the printer port from the PC are terminated by 4.7 k Ω pull-up resistors (R5) and buffered by the 74HC245 bus driver IC1. This IC carries the signals that switch voltages within the programmer so it is important to ensure that its supply voltage is derived from a separate (unswitched) 5 V regulator (IC4). Serial data for the PROM also passes through IC1 and IC3. The supply for IC3 comes from the switched V_{CC} . This ensures that voltage thresholds of the PROM signals correspond to V_{CC} and that the signals DATA, CLK, \overline{OE} and \overline{CE} (and the acknowledge signals ACK and PE) will only be driven when the supply voltage V_{CC} is switched on. To be absolutely sure that V_{CC} cannot be pulled high via the input protection diodes of IC3, transistor T5 switches the supply to ground via the 1 k Ω resistor when V_{CC} is switched off.

The programmer can handle both 3.3 V and 5 V PROMs so it's important to use an **HC type** of driver for IC3. The LS or ALS family of devices do not provide suitable output signal

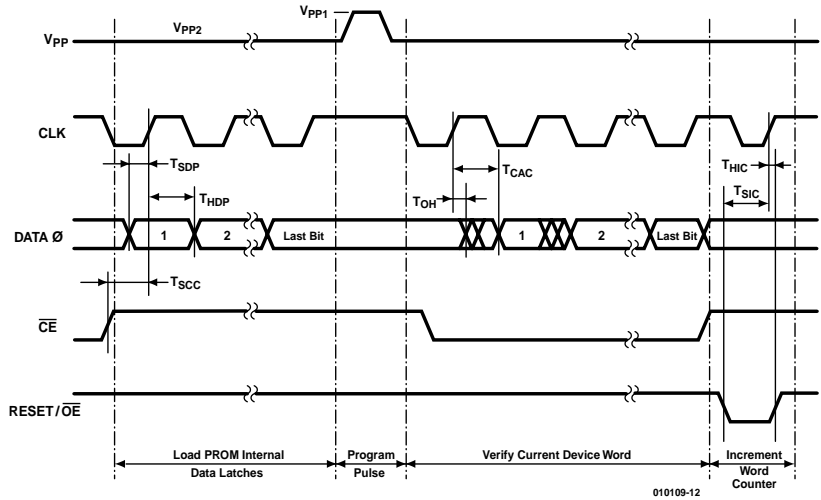


Figure 1. Programming cycle timing diagram for the XC17VXX type PROM.

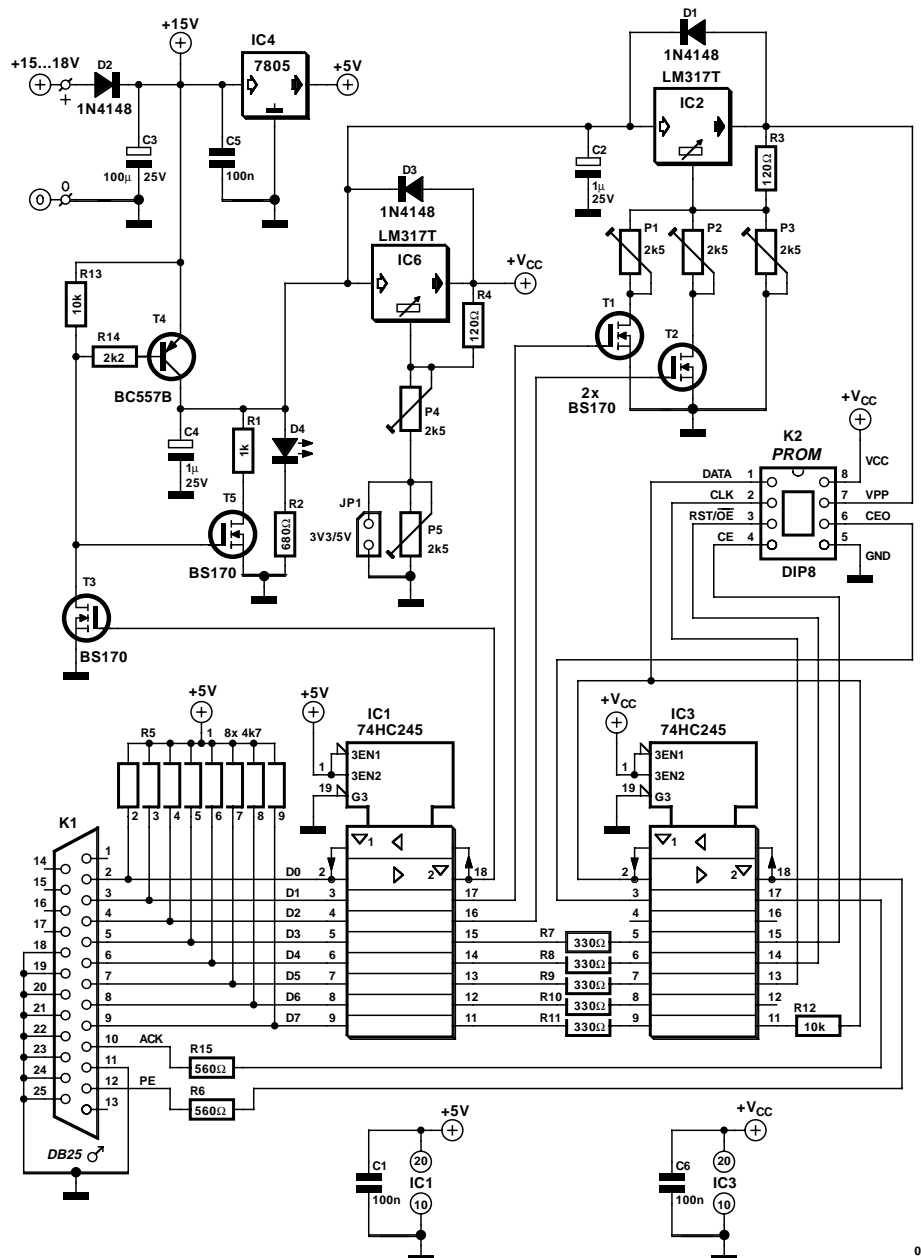


Figure 2. The circuit diagram.

COMPONENTS LIST

Resistors:

- R1 = 1k Ω
- R2 = 680 Ω
- R3,R4 = 120 Ω
- R5 = 8-way 4k Ω 7 SIL array
- R6,R15 = 560 Ω
- R7-R11 = 330 Ω
- R12,R13 = 10k Ω
- R14 = 2k Ω
- P1-P5 = 2k Ω 5 preset

Capacitors:

- C1,C5,C6 = 100nF
- C2,C4 = 1 μ F 25V radial
- C3 = 100 μ F 25V radial

Semiconductors:

- D1,D2,D3 = 1N4148
- D4 = LED, red
- IC1,IC3 = 74HC245
- IC2,IC6 = LM317T
- IC4 = 7805
- T1,T2,T3,T5 = BS170
- T4 = BC557B

Miscellaneous:

- JP1 = 2-way SIL header with jumper
- K1 = 25-way sub-D plug (male), PCB mount
- K2 = 8-way DIL socket with turned pins PCB, order code **010109-1**, available from The PCBShop
- Disk, project software, order code **010109-11** or Free Download

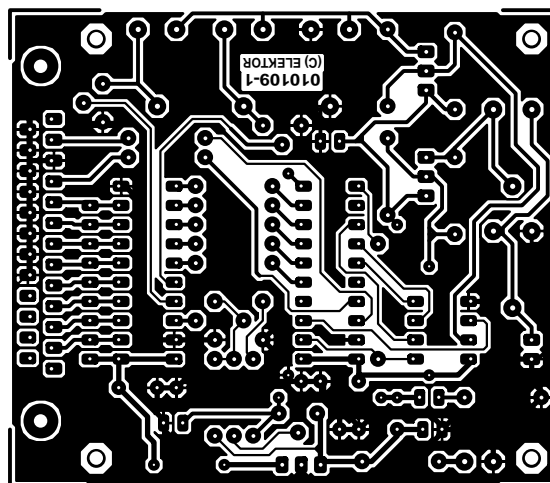
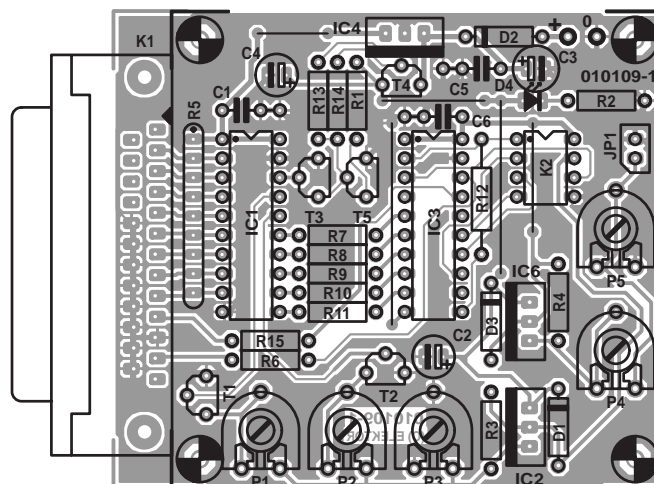


Figure 3. The Xilinx programmer uses a single sided PCB.

levels and the HCT family cannot operate at the necessary supply voltages. HC devices can operate with a supply of between 2 V and 6 V. The series resistors in the data inputs of IC3 limit the input current when $V_{cc} < 5$ V.

The DATA pin of the PROM can be configured as either an input or output. Resistor R12 is used to effectively increase the output impedance of IC3 pin 11, this ensures that whenever the PROM drives data out from pin 1 it will always be available as an input to the PC via line PE irrespective of the state of the output of IC3 pin 11. It is only when writing to the PROM (when the DATA pin will be an input and therefore high impedance) that the output of IC3 pin 11 (write data) will appear on the DATA pin. An important parameter of the data transfer is the slope of the signal edges of the PE connection to the PC. Resistor R6 is used to reduce the effects of reflections in the cable that may otherwise corrupt the signal.

The circuit requires a single external input voltage in the range of 15 to 18 V; all other voltages required for programming are

derived on-board. To be on the safe side all PROM voltages are switched by signals via the printer port.

The programmer can use PROMs with either a 5 V or 3.3 V supply voltage (V_{cc}), jumper JP1 selects between these two supply levels and gives a simple visual indication of the supply voltage selected: with JP1 fitted the supply voltage is 3.3 V, with JP1 not fitted the supply voltage will be 5 V. Pin 7 (V_{pp}) requires three different voltage levels depending on the operation that the PROM is performing: V_{pp1} for programming, V_{pp-nom} for reading the data and V_{pp2} for verification of the writing procedure. Transistors T1 and T2 switch these three voltage levels from data signals D1 and D2. With V_{cc} switched on and D1 and D2 low the level on this pin will be V_{pp1} . The programmer voltage sequencing

is given in the software. The program also indicates the correct voltage levels necessary depending on the type of PROM selected so it's not necessary to wade through piles of data sheets for this information. The LED indicates that reading and programming voltages are available.

Software features

The programmer software is written in Turbo Pascal and runs under DOS using command line options. Among the software downloads for this article a free version of DOS is included, this enables a boot disk to be made so that the programmer software will run even if you do not have DOS available on your machine. Using the command **spb /?** gives an overview of all the available options such as selection of chip type and

the type of access required of the PROM (eg reading, writing or reset bit reading etc). The program always asks if the chip is plugged into the socket before it attempts to switch the supply voltage to the PROM. The program checks if the PROM in the socket corresponds with type of chip selected and also documents the programming process.

The program options are described below in the same order that they are used in the program:

/?

Overview of the program options.

/d?

The printer port where '?' is the number of the LPT port connected to the programmer.

/av

This command is used to setup the different programming voltages required for the specified PROM. An oscilloscope or DVM can be used to test voltage levels. The program indicates the required voltage and on which pin of the PROM socket the voltage should be measured. It also indicates which preset needs to be adjusted. The chip name is only used here for the sake of appearance.

/m

This function reads the chips MfG code. A chip name must be entered here. When the MfG code is correct then the message Chip identified will be displayed; if not the MfG code read from the chip will be displayed. If it is necessary to enter a new chip into the chip information file (chip-info.dat), all the displayed values must first be converted to hexadecimal.

All further functions are specific to each chip type and will only be carried out if the MfG code read from the chip corresponds to the selected chip type. The MfG code is always tested first.

/rp

Reads the polarity of the reset bit.

/bc

Blank Check, this tests if all the PROM bits are blank (=1). The programming specification suggests that only a few bits need to be

checked to determine if the chip is blank but to be sure the program performs a blank check on the entire PROM, the extra time needed is not excessively long.

/b

Burn, the data held in the brenndat.bin file is programmed into the PROM, this operation will not program the polarity of the reset signal. The file name is stored as the constant BrenndatenFileName and can be changed in the program at this point. The size of the data in the file must always be smaller or equal to the available space in the PROM. If the file size is bigger than the PROM size given in 'Size' an error message is generated and the burning process will not start.

/r

This command reads all the data in the PROM and stores it in the file data.bin. This feature will execute differently depending on whether the reset bit has already been programmed. The reset polarity should always be read before this command is used.

/cp

Changes the reset polarity. First the polarity of the reset is tested using /rp and if necessary the reset polarity is changed. Lastly the reset status will be read and displayed.

Two examples

In order to read a XC17S10 type of PROM connected to port LPT1 the following command line is used: spb /r s10 /d1

Where s10 specifies the PROM type. Using the command spb /av /cp /b s200a /d2 indicates that the PROM type has changed and the programming voltages will be set up and adjusted before the PROM is inserted into the socket. The reset polarity will then be changed and Data will be programmed into the XC17S200A via the LPT2 printer port.

In and out

In order to be able to switch all the voltages and signals in the programmer hardware each bit of the printer port byte must be individually programmable. In the software the state

of the eight printer port output bits is held in the eight bit wide global variable dat. As an example if we wanted to switch on the supply voltage to the programmer we would need to set bit D0 (connector K1 pin 2) to a 1. This could be accomplished with the line:

```
dat := dat or $01
```

This will set the bit in software and then with an output instruction to the printer port this new value of D0 will be output to the programmer leaving bits D1 to D7 unchanged. To turn off the supply voltage we just need to do the reverse procedure:

```
dat := dat and $fe
```

or alternatively:

```
dat := dat and (not $01)
```

And again write it to the printer port. The hexadecimal value of each bit of the LPT port (and its inverse hex value) is given below:

\$01 (\$FE)	Vpp - Relay (REL)
\$02 (\$FD)	Vpp-nom
\$04 (\$FB)	Vpp2
\$08 (\$F7)	\overline{CE}
\$10 (\$EF)	RESET/ \overline{OE}
\$20 (\$DF)	CLK
\$80 (\$7F)	DATA

Chipinfo.dat

All the data relating to the chip are stored in this file. The information required by the programmer is:

Mm This byte in the identification code identifies the chip manufacturer (Xilinx = \$C9).

IDCLK This is the number of clock pulses needed to read out the manufacturers code.

ID is the second part of the MfG code indicating the chip series and type.

Size is the size of the PROM in bits.

RSTClk This is the number of clock pulses needed to program the Reset bit.

Vcc gives the PROM operating voltage.

Vpp1 gives the PROM programming voltage.

Vpp2 This voltage on pin 7 used to test/fix the data written into the PROM. This voltage level is only slightly higher than the supply voltage and is not implemented at present since verification at this slightly raised voltage is not essential.

Vppnom The voltage on the pin V_{PP} during reading. It will be at least as high as the chips supply voltage V_{CC}.

Bugfix

The program is written in Turbo Pascal and uses a system unit called *Crt*, this unit includes a bug which causes a run time error in Windows if the PC's processor is a Pentium type running faster than about 200MHz. The program runs without problem in DOS.

A patch is included in the software package **010109-11** for this article. The patch program (TPPATCH.EXE) is able to locate and correct the failure in this unit. If you need to make any changes to the source code for this project it is therefore important to run this patch program once you have recompiled *spb.exe*.

Name This is an abbreviated name identifying the type of PROM used in the command line to call the program. The same name must not be used for PROMs from other manufacturers even if the PROM has the same memory capacity.

The four voltages V_{CC} , V_{pp1} , V_{pp2} and V_{ppnom} are used to simplify the voltage settings when different types of PROMs are used. The levels are displayed by the /av function. The file can be expanded to incorporate other serial PROMs that have compatible programming algorithms. The maximum number of entries is given by the *maxdats* constant defined in *spb.pas* and will need to be changed if new devices are added. All the data in *chipinfo.dat* is taken from [2].

Reading the MfG code

Each PROM contains its manufacturer's identification code that enables it to be recognised by the programmer. The code consists of two bytes: the first byte identifies the manufacturer (Xilinx uses the code \$C9). The second byte comprises four bits indicating the density of the PROM and four bits for information on the programming algorithm. This is contained in the second column of *chipinfo.dat*.

The PROMs manufacturer code is read in the *ReadMfGCode* procedure. The PROM must first be set to program mode with $\overline{CE} = 1$ and $\overline{OE} = 0$. Next a defined number of clock pulses is sent to the PROM these are described as the N_{IDCLK} or also the *IDCLK* (see the first column of the *chipinfo.dat*). Now with $\overline{OE} = 1$ and $\overline{CE} = 0$ the MSB of the MfG-Code is output on the *DATA* pin. The code is now read by the *LeseByteMfG* function as each bit is shifted out by the clock signal.

The Reset Bit

The polarity of the serial PROMs reset signal can be programmed. When the PROM is used together with a Xilinx FPGA it is useful to be able to reset the PROM from the INIT signal from the FPLA. This means that the polarity of the PROM reset pin needs to be changed so that $RESET = 0$. Changing the polarity of the reset signal is a one-way operation so once the polarity has been programmed to negative reset it cannot be changed back. For trials it is best to test the chip with an unprogrammed reset ($RESET = 1$).

Programming

The Burning procedure switches the PROM into programming mode. Both \overline{CE} and \overline{OE} must be high for at least two positive clock edges and the voltage on V_{pp} must be at V_{pp1} [1]. In this mode the manufacturer and chip information can be read and also data can be written to the PROM. The data is shifted serially into the PROM in 64 bit chunks and latched by a V_{pp1} pulse on pin V_{pp} . The XC17S05D Xilinx PROM is the only type with 32 bit long latches; it is not widely used and is not supported by this programmer. According to the specification after programming the voltage level on V_{pp} should be set to V_{pp2} but this is not implemented in the program. Lastly the programmer switches out of program mode and both supply voltages V_{CC} and V_{pp} are switched off.

Build and operation

The single-sided PCB has a board area of 72.63 mm², this gives enough space for the components to be mounted quite easily. The five wire links need to be fitted first (one of these runs under the 8 pin DIP programming socket).

Once you are sure that all the polarised components have been fitted the correct way round (including the SIL resistor network) and double-checked your soldering the circuit can be powered up. The first test should take place without a PROM fitted to the programming socket. Connect 15 V to 18 V between the supply solder pins on the PCB. Diode

D2 ensures that no damage occurs if you accidentally mix up the power supply polarity. First check that 5 V is available on pin 20 of IC1, LED D4 should also light to indicate that the supply voltage is available.

The software can now be installed and run to drive the programmer. Jumper JP1 allows the supply voltage of the PROM to be quickly changed between 5 V and 3.3 V depending on the type of PROM used. This supply can also be adjusted to suit PROMs using other supply voltages. Xilinx are currently offering three PROMs families each with a different supply voltage. The supply voltage data for the PROMs is stored in the *Chipinfo.dat* file and can be read using the /av option so the voltage levels can be quickly adjusted.

After installation the PCB can be plugged into the printer socket of the PC. The LED will go out and commands can be entered to the programmer software. For example using the command line *spb /av /d1 s10* calls the program where /av indicates the voltage test and /d1 is for printer port LPT1 (d2 would be entered here if LPT2 was used) and s10 indicates the PROM type.

Next it is necessary to adjust the two-level supply voltage to the PROM. Firstly fit jumper JP1 and with a DVM connected to the output of IC6 (V_{CC}) adjust the preset P4 until V_{CC} reaches 3.3 V, once this has been achieved remove JP1 and adjust preset P5 until V_{CC} is 5 V.

Finally the program indicates adjustment of the programming voltages on pin 7. These settings are best made when you need to change to a different type of PROM.

(010109-1)

References:

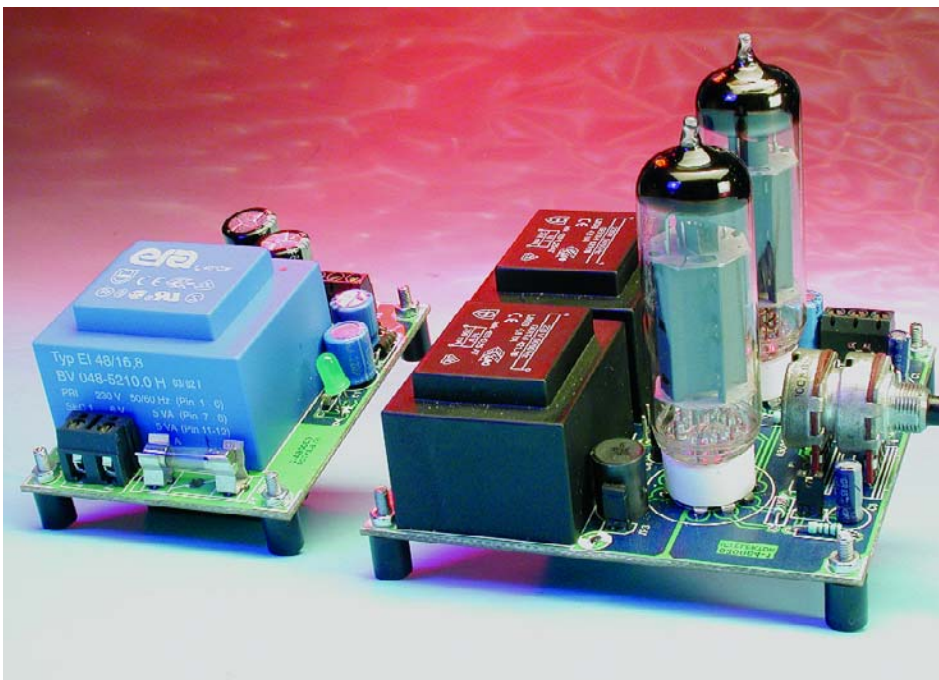
- [1] Programmer Specification
www.xilinx.com/support/programr/specs.htm
- [2] PROM Specification
www.xilinx.com/isp/csapps.htm

Valve Headphone Amplifier

EL84 (6BQ5) at an anode voltage of 40 V

Design by B. Kainka

It is not just power amplifiers that can deliver the much sought-after warmth of the 'valve sound' — headphone amplifiers can too. What is special about this project is that it is achieved using a safe anode voltage of just 40 V.



Heated debates rage over whether, and how, a valve amplifier sounds different from (or perhaps better than) a modern semiconductor amplifier. There certainly are reasons why a valve amplifier might sound different. First there is the particular form of the characteristic curve (I_a plotted against U_g), whose gentle curve inevitably gives rise to increased distortion as the drive level is increased,

especially when negative feedback is not employed. A modern semiconductor amplifier, on the other hand, almost always uses plenty of negative feedback in order to keep distortion to a minimum. However, the human ear itself is non-linear at higher sound levels, and so the 'natural' distortions of a valve can give

the impression of a higher volume. Also, a transistor output stage suddenly produces extreme levels of distortion when overdriven, whereas a valve output stage clips more gently.

A second, decisive, factor is the extremely high internal impedance of a valve. The connected audio transducer is therefore practically undamped, whereas in the case of an extremely low-impedance transistor output stage all the self-resonances of a loudspeaker or headphones are strongly damped. This gives a flatter frequency response but simultaneously suppresses the particular character of the audio transducer. This is one of the reasons why the sound of older valve equipment is so often found preferable. At the same time it is also behind a technical problem encountered in the design of valve amplifiers, since a transformer is often required to match the low impedance of the audio transducer.

As mentioned in the article in the September 2003 issue of *Elektor Electronics*, the most important feature of this headphone amplifier is its safety, since it operates with a low anode voltage. The amplifier requires a line-level input of 1 V peak-to-peak and provides plenty of output power.

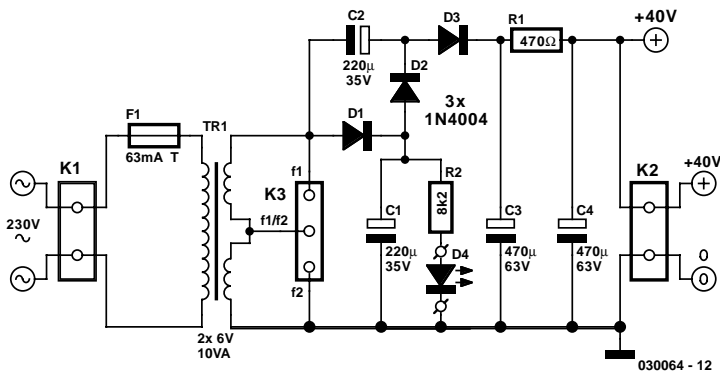


Figure 1. A simple power supply.



EL84 in strange surroundings

The EL84 (or its US equivalent the 6BQ5), a 5 W power output valve usually found in radios, might seem a surprising choice here. Does it really make sense to operate this valve from just 40 V instead of the 250 V suggested in the data sheet? It does: the anode voltage of 40 V is safe and entirely adequate for our purposes here. We do get a lower transconductance and a considerably lower anode current (of around 5 mA), but this does not do any harm to the valve. The anode power dissipation will thus be some 200 mW, and so there should be more than enough output power for driving headphones.

The EL84/6BQ5 is also relatively inexpensive and readily available: it is still in production, as it is still used in many medium power hi-fi amplifiers. Of course, an EL34 or an EL504

could also be used, although these devices are respectively more expensive and harder to obtain.

A further reason for choosing the EL84/6BQ5 is that it makes constructing the power supply more straightforward. The valve requires a heater voltage of 6.3 V and a heater current of 0.7 A, for a heater power of about 4.5 W. A 10 VA encapsulated mains transformer with two 6 V secondaries makes a suitable supply. In theory the two heaters can be wired in series and a 12 V transformer used, but it is found that in practice the heater voltages are not then equal. An arrangement like that in **Figure 1** is better, where each valve is supplied from its own transformer winding. The two windings are nevertheless in series, and it is straightforward to generate a DC supply at about 40 V for the anodes from this 12 V AC supply: here we use a tripler circuit.

Amplifiers with and without negative feedback

The amplifier circuit in **Figure 2** shows a fairly typical class A output stage, as used in loudspeaker amplifiers. The anode currents do not flow through the output transformers, since the magnetisation caused by the DC component would lead to distortion. The audio frequency signal is taken via a volume control potentiometer to grid 1 (G1) of the valve. A 1 kΩ series resistor suppresses high frequency oscillations. The screen grid (grid 2) is at supply potential, while the suppressor grid (grid 3) is connected to the cathode. The output signal at the anode is coupled to the audio transformer via an electrolytic capacitor. Alternatively, high-impedance headphones may be connected directly to the coupling capacitors.

Compared to a triode, a pentode has a very high internal impedance. This is extremely helpful in designing the amplifier and allows inexpensive mains transformers to be used at the output.

The voltage drop across the cathode resistor provides for a negative grid bias voltage relative to the cathode. The operating point of the valve is set by this resistor. A value of 100 Ω provides a grid voltage of 0.5 V with an anode current of 5 mA. At the same time, the cathode resistor provides a certain amount of negative feedback and hence a reduction in distortion in the valve, without reducing the internal impedance of the amplifier.

A particular feature of this circuit is that the negative feedback can be disabled by fitting jumpers JP1 and JP2, bringing the parallel 100 µF electrolytic capacitors into play. This affects the sound and the output power of the amplifier.

The two characteristic curves — at an anode voltage of 40 V and at 250 V — exhibit a gentle curve and are very similar to one another. It is precisely here that the secret of 'valve sound' lies. If no attempt is made to straighten the characteristic curve using negative feedback, a certain distortion to the audio signal inevitably arises. These comprise harmonics and mixing products, chiefly odd multiples of the fundamental frequency, which are pleasant to the human ear. Since the ear does not work linearly at high volumes, a live concert at 100 dB(A) sounds better than a recording. With a valve amplifier the sound is also pleasant, more 'full', at lower volumes.

In the quiescent state, with $U_g = -0.5$ V, no grid current flows. When it is biased to 0 V the peak grid current can be up to 20 µA, which flows through the 100 kΩ volume control potentiometer acting as a grid leak resis-

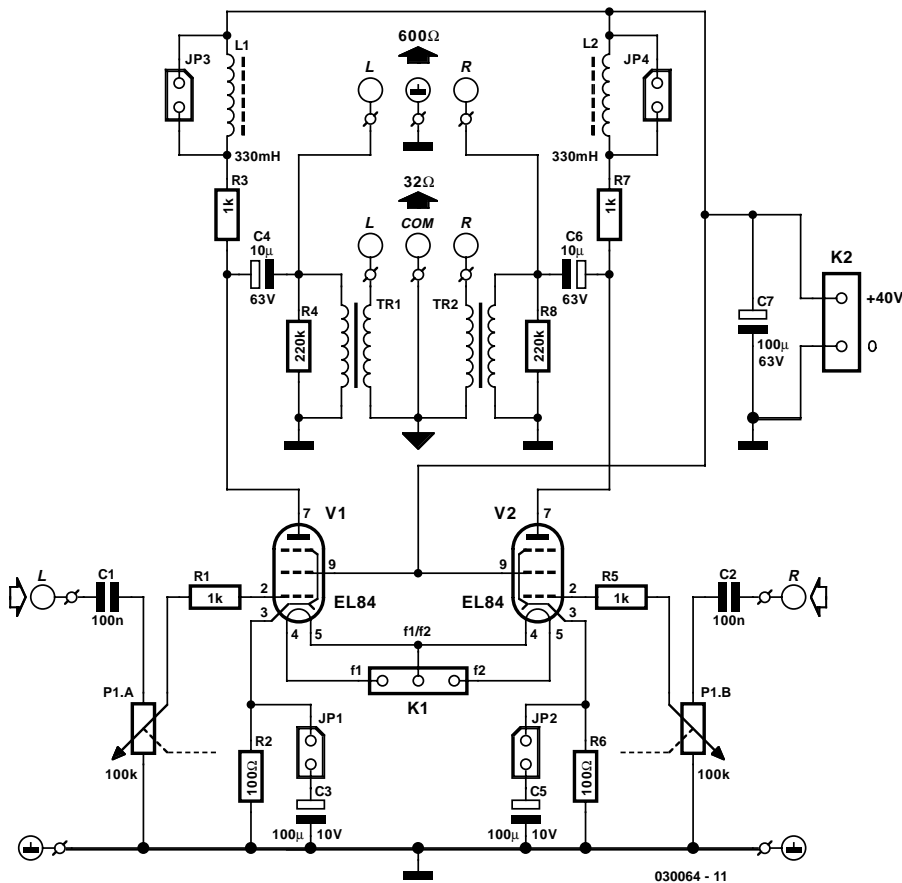


Figure 2. The amplifier with switchable negative feedback.

tor. A conventional power amplifier with a higher anode voltage would use a value of 100 kΩ to 1 MΩ here. At lower anode voltages this value is critical. Too high a grid leak resistance would result in an additional drop in grid voltage when a grid current flows. On the other hand, the input impedance of the amplifier must not be allowed to go too low, or it may not be compatible with the line output of ordinary hi-fi equipment. A 100 kΩ potentiometer provides a suitable compromise.

The subtleties of transformer selection

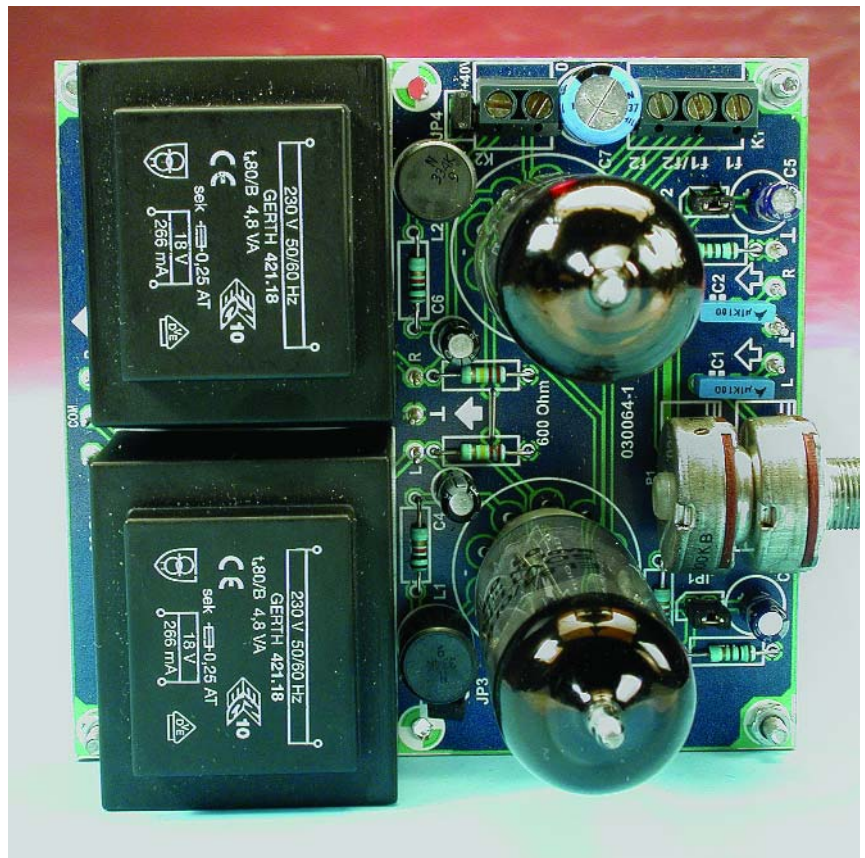
Another key to constructing a successful amplifier is the selection of the output transformer. A genuine audio transformer would have to be specially made for this application and would therefore be very expensive. For that reason, we shall use a readily-available mains transformer. Excellent results can be achieved as long as the right type is used. A mains transformer can handle frequencies above 50 Hz if it has a suitable core size and the windings have the right inductance and DC resistance.

Inductance

The output impedance of the valve is closely approximated by $R_a = U_a / I_a$. According to the EL84 data sheet, the valve should be operated at 250 V and 48 mA, implying an external impedance of around 5 kΩ. The data sheet actually recommends between 4.5 kΩ and 5.2 kΩ. At 40 V and 5 mA we obtain 8 kΩ. If anything, the value chosen should be lower rather than higher than this. Too low an external resistance leads to a lower AC component of anode voltage at the extremes of anode current. Since this means that the anode voltage is less likely to reach zero, the result is a lower likelihood of distortion.

Consider a transformer with a turns ratio of 230 V:18 V = 12.8:1. The impedance ratio is to a first approximation $12.8^2:1 = 164:1$. Thus a headphone impedance of 32 Ω is transformed into an external impedance of 5240 Ω. An 18 V mains transformer would therefore appear to be suitable.

This argument is valid only when the transformer is directly in the anode circuit. In practice the 1 kΩ anode resistors reduce the impedance of the circuit, and so for optimal



matching a lower turns ratio is preferable. The additional coils L1 and L2 increase the impedance at higher frequencies, which, together

with the transformers used, gives an overall flat frequency response. When high-impedance headphones are being used these coils should be

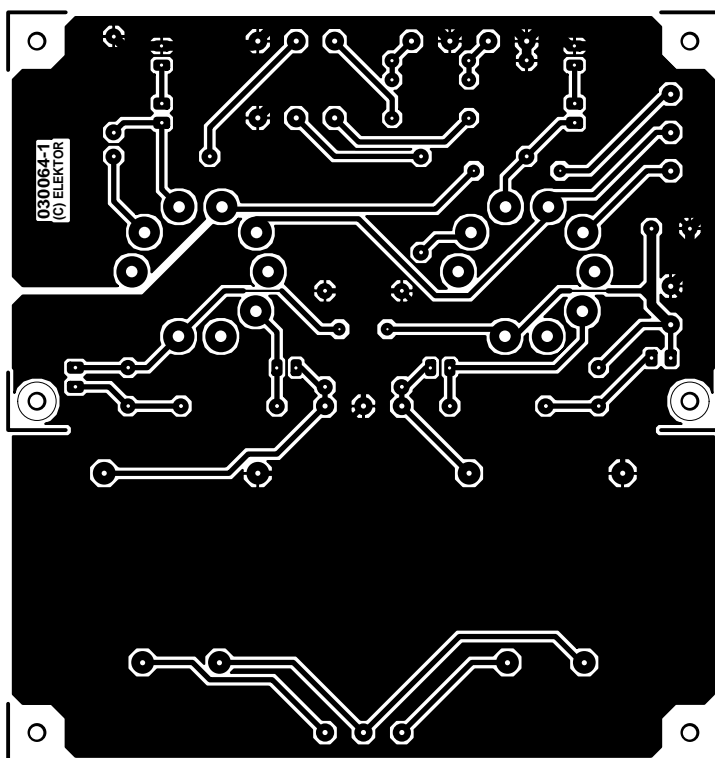
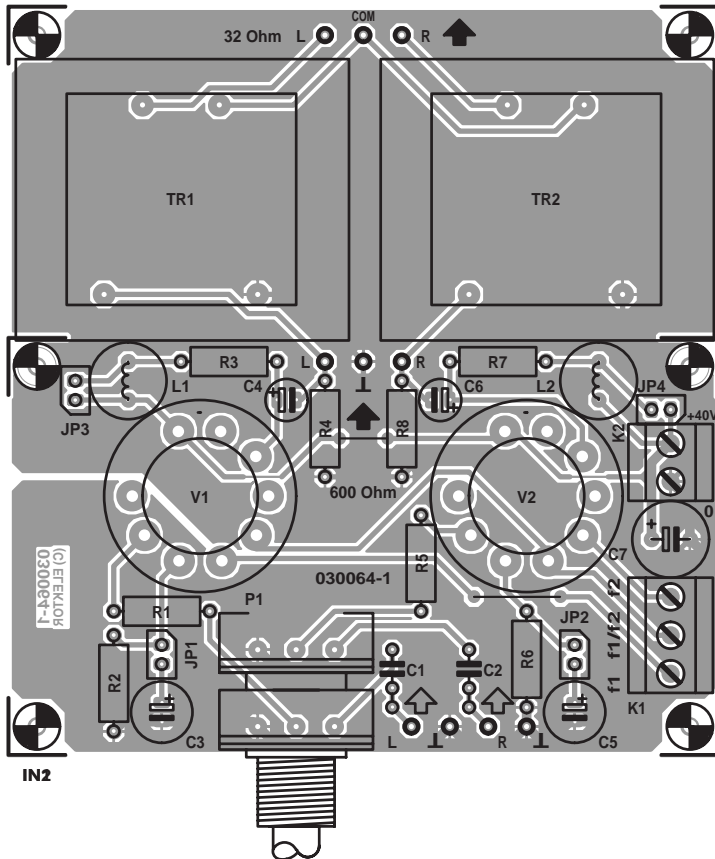
shorted by fitting jumpers JP3 and JP4.

Core size

In selecting a suitable core size for the transformer (which corresponds to rated power) the general principle is that a smaller transformer has a higher inductance and so can operate at lower frequencies. Also, smaller transformers have thinner windings and therefore greater resistance. So how large should the transformer be in order to guarantee the desired frequency response? The second consideration is the unavoidable leakage inductance of the transformer. This can be thought of as being in series with the primary winding and so forms a low-pass filter, which, in extreme circumstances, can impair performance at high frequencies.

DC resistance

As well as the above, the DC resistance of the windings and the exact turns ratio are important. Manufacturers often specify the open-circuit voltage ratio of small mains transformers. A transformer with a secondary voltage of 10 V at rated load might, for example, deliver 13 V with no load, giving a load regulation of 30%. This must be taken into account



COMPONENTS LIST

Resistors:

- R1,R3,R5,R7 = 1kΩ
- R2,R6 = 100Ω
- R4,R8 = 220kΩ
- P1 = 100kΩ stereo logarithmic potentiometer

Capacitors:

- C1,C2 = 100nF
- C3,C5 = 100μF 10V radial
- C4,C6 = 10μF 63V radial
- C7 = 100μF 63V radial

Valves:

- B1,B2 = EL84 or 6BQ5 with noval socket (18mm diam.)

Inductors:

- L1,L2 = 330mH choke (Toko 10RBH 239LY334K)

Miscellaneous:

- JP1-JP4 = 2-way pinheader with jumper
- K1 = 3-way PCB terminal block, 5mm lead pitch
- K2 = 2-way PCB terminal block, 5mm lead pitch
- TR1,TR2 = mains transformer 18V/4.8VA (Gerth # 421.18)
- PCBs, order code **030064-1** and **030064-2**

Figure 3. Printed circuit board layout with separable 32 Ω section.

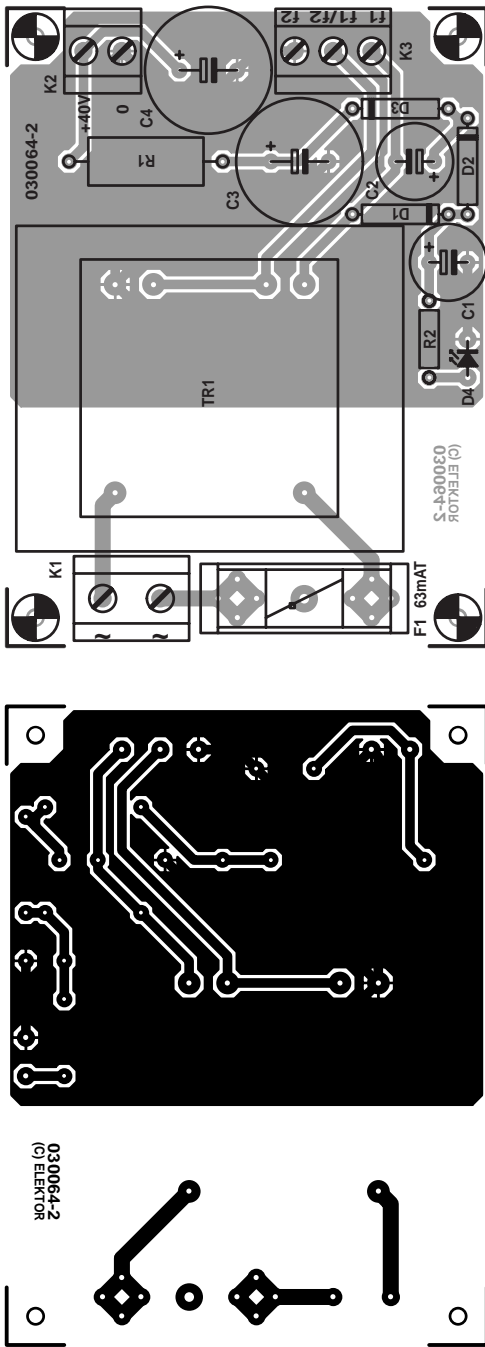


Figure 4. Power supply printed circuit board.

when calculating the actual turns ratio. A good choice is a transformer with an EI42 core and rated at about 5 VA. The Gerth 4200 range includes a suitable type (available for example from Reichelt Elektronik, Germany, order code 421.18-1). It is not too large, and, at about two pounds, not too expensive either. It will guarantee good sound quality with 32 Ω headphones.

High or low impedance?

A frequency response from 30 Hz to 20 kHz

Characteristic curves

The characteristic curve in Figure A shows how the valve behaves at the reduced anode voltage of 40 V. Comparison with the curve from the data sheet (Figure B) shows similar behaviour, although only at rather higher currents. This means that we can use the EL84 in the same way as it is used in a valve radio, albeit with lower output power. Also, when using an anode voltage of 40 V, the grid bias voltage must be reduced.

Comparing the two characteristic curves shows that its position shifts at lower anode voltages with lower anode currents and lower grid bias voltages. The optimal operating point lies at around $U_{g1} = -0.5$ V and $I_a = 5$ mA. At full drive with an audio frequency signal of 1 V peak-to-peak the valve is then driven between -1 V and 0 V, giving an anode current varying between 3 mA and 8 mA.

When driven to around 0 V or slightly more, the grid current also becomes important. For this reason we also need to consider the grid current characteristic (Figure C), which in turn is somewhat dependent on the anode voltage. A higher anode voltage reduces the grid current, since free electrons are drawn more towards the anode, and fewer land on the grid.

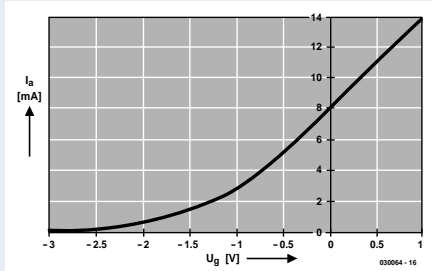


Figure A. Characteristic curve for the EL84 at $U_a = U_{g2} = 40$ V.

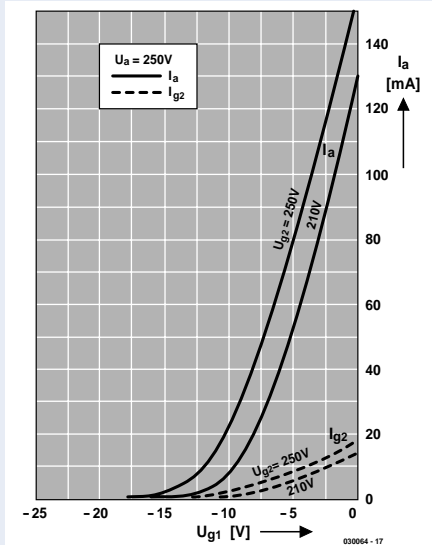


Figure B. Characteristic curve at 250 V.

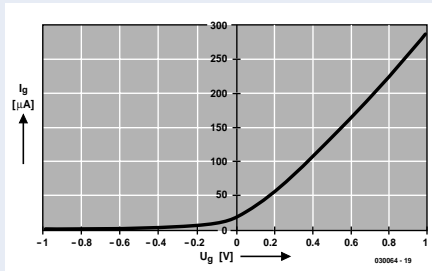


Figure C. Grid current up to $U_g = +1$ V.

into low-impedance headphones is very good when it is taken into account that it is being achieved using an inexpensive mains transformer. The alternative would be a very expensive custom device using layered windings to reduce leakage inductance. The audio transformers used in professional valve output stages are large and expensive.

An alternative to these specialist transformers is to experiment with types designed to be connected to a

600 Ω output. An interesting possibility would be to try ordinary 100 V audio line transformers (such as Conrad Electronics order code 516104-77), which have multiple taps on the primary and secondary allowing the matching to be adjusted.

The effort in finding a suitable transformer for the output stage is worthwhile: 32 Ω headphones deliver a pleasant and warm sound, even though the frequency response is not as linear as in a transformerless

Cutoff frequencies

The equivalent circuit of the transformer in **Figure A** illustrates the connection of 32 Ω headphones. The wire resistance of the secondary can be considered as 10 Ω in series with the load. There is thus a total of 42 Ω across the output. The actual turns ratio of 9.8 : 1 gives an impedance ratio of 9.8² : 1 = 96 : 1. The primary thus has an impedance of 4070 Ω + 875 Ω = 4945 Ω, say 5 kΩ. This value fits well with the theoretical value of 8 kΩ, which, in the interests of a pleasant sound, should be regarded as an upper limit. The low frequency response is limited by the high-pass filter comprising this load resistance and the parallel inductance of 14 H (**Figure B**). The -3 dB point of this filter can be computed as 56 Hz. In practice, however, we measure a lower cutoff frequency of around 30 Hz. Presumably we can attribute this difference between theory and practice to the difficulties inherent in measuring the inductance of the transformer.

The upper cutoff frequency arises from the low-pass filter comprising the load resistance and the leakage inductance (**Figure C**). Using figures of 4945 Ω and 0.5 H, we calculate the cutoff frequency, rather disappointingly, to be 1574 Hz. A value in this range is indeed measured if the transformer is connected to a low impedance signal source. Fortunately the dynamic impedance of the valve is at least ten times greater than the theoretical load resistance (here about 8 kΩ). Since, thanks to coils L1 and L2, the signal source has an impedance of over 80 kΩ at the higher end of the frequency range, the theoretically calculated upper cutoff frequency increases to over 25 kHz. The leakage inductance also comes into play: at higher signal levels the valve more readily goes into clipping. In practice this is often not a significant effect, since the distortion products lie outside the audio range.

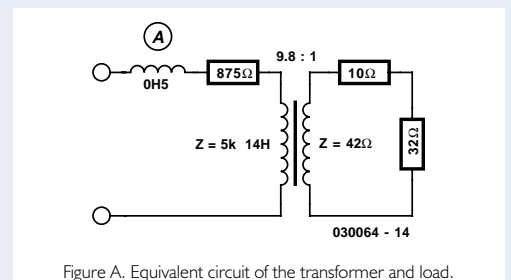


Figure A. Equivalent circuit of the transformer and load.

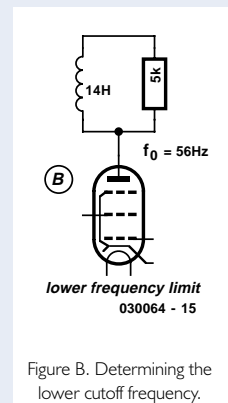


Figure B. Determining the lower cutoff frequency.

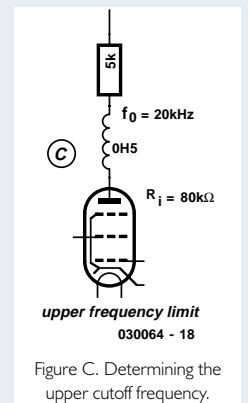


Figure C. Determining the upper cutoff frequency.

amplifier. None of these problems arises when using high-impedance headphones and capacitor coupling.

Do not forget to compare the sound with and without the cathode capacitors. The choice here is between lower distortion and more 'valve sound'. A little negative feedback will do no harm when listening to classical music in all its transparent glory; but rock music definitely calls for a bit of valve-style distortion.

Construction

The printed circuit board in **Figure 3** includes a section providing 32 Ω matching that can be separated from the rest of the circuit. There are no particular points to note about construction: there are only two wire links (near R4 and R6). If the Toko chokes cannot be obtained, types from other manufacturers can be substituted. Since coils over 100 mH are rare, it is also possible to use smaller values in series. Alternatives include:

Neosid BS75 (part number 00612436, 100 mH, 480 Ω, I_{max} = 5 mA, radial)

Fastron XHBC (part number XHBC-104J-01, 100 mH, 245 Ω, I_{sat} = 60 mA, axial)

Epcos B82144-A (part number B82144-A2107-J, 100 mH, 420 Ω, I_r

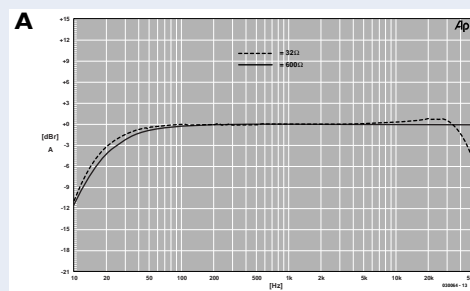
= 20 mA, axial)

Of course, you can also wind the coils yourself using an RM8 former and an N67 core.

If more output power is required, transformers with a higher secondary voltage can be tried. In this case, L1 and L2 will probably need to be adjusted.

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Frequency response



The measurement illustrated in **Figure A** shows the relative frequency responses of the amplifier with a 600 Ω load at the output (solid line) and with a 32 Ω load (dotted line). The curves were obtained with different anode voltages.

Input impedance	100 kΩ	
Sensitivity	600 Ω, 1 mW, JP1/JP2 open	620 mV (THD = 4.5 %)
	600 Ω, 1 mW, JP1/JP2 closed	370 mV (THD = 7.4 %)
	33 Ω, 1 mW, JP1/JP2 open	0.94 mV (THD = 7.5 %)
	33 Ω, 1 mW, JP1/JP2 closed	0.59 V (THD = 9.9 %)
Signal to noise ratio (1 mW, JP1/JP2 open)	600 Ω	>62 dB (B = 22 kHz lin.)
	33 Ω	>88 dB(A)
		>65 dB (B = 22 kHz lin.)
THD+N (1 kHz, B = 80 kHz, JP1/JP2 open)	600 Ω/1 mW	4.5 %
	600 Ω/0.1 mW	1.1 %
	33 Ω/1 mW	7.5 %
	33 Ω/0.1 mW	3.5 %
Bandwidth	23 Hz to >200 kHz (600 Ω)	
	20 Hz to 45 kHz (33 Ω)	

The P89LPC900

Part I: the architecture

By M. Stottelaar

Eurodis Texim

Under the moniker of P89LPC900, Philips have introduced an entirely new family of 80C51 successors that have a number of pins ranging from 8 to 80 and are all based on an (in-system-programmable) flash memory. An important characteristic of the controller is that despite a large number of new features it is still fully compatible with the 80C51.

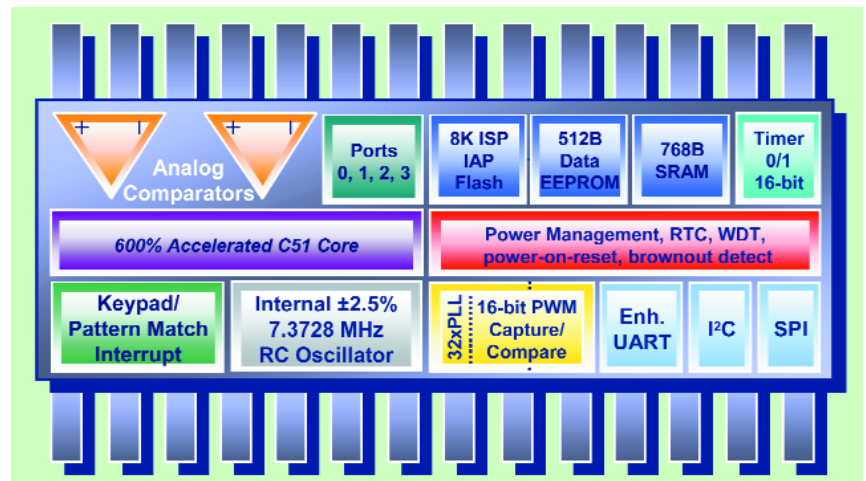
That the 80C51-based controllers are extremely popular is nothing new, certainly when considering the large number of designs that have appeared in this magazine over the years. The reason may well be the fact that the tools (both hardware and software) that are available for this controller are very affordable and there is an enormous amount of information readily available. In addition, a very active forum provides answers to many questions.

Philips, one of the biggest players in the 80C51 field has, with the part number P89LPC900, recently breathed new life into this architecture and improved the controller family to the point that its lifespan has been extended for many more years.

In this article we will describe the architecture of the P89LPC900. In a follow-up instalment we will publish a simple programmer/I/O-board.

The heart of the controller

One of the most significant features of the P89LPC900 is that the core now requires only **two clock cycles per instruction**. 8051 experts will already know that this used to be 12 or 6 cycles until now. But to be fair, it needs to be mentioned that the maximum clock frequency has been reduced to 12 MHz (167 nsec per instruction). The net improvement is still more than a factor of two, compared with previous solutions that used to run at 33 MHz. The improved performance at lower crystal frequencies is beneficial to power consump-



tion and EMI. The P89LPC900 is therefore very appropriate for low-energy designs such as battery-powered devices. On average, the current consumption per MHz is not much lower than before, but you do get a six times improvement in performance. In practice, this means that the crystal frequency can be drastically lowered. The lowest possible current consumption, by the way, is only 1 μ A.

The clock

Most of the existing 80C51 controllers have a relatively simple clock distribution circuit. A crystal, res-

onator or external clock signal is connected directly to the CPU and also to the peripherals. In contrast, the P89LPC900 has a much more elaborate mechanism, which allows the user to determine where the clock source comes from and how it is to be presented to the CPU and peripherals. A standard, and in the context of microcontrollers, very accurate internal RC oscillator (2.5% over the entire temperature range) will in many cases suffice as the clock source. Even a UART will generally operate very well with this kind of accuracy. In the event that the requirements are more stringent, the processor offers the option of mea-

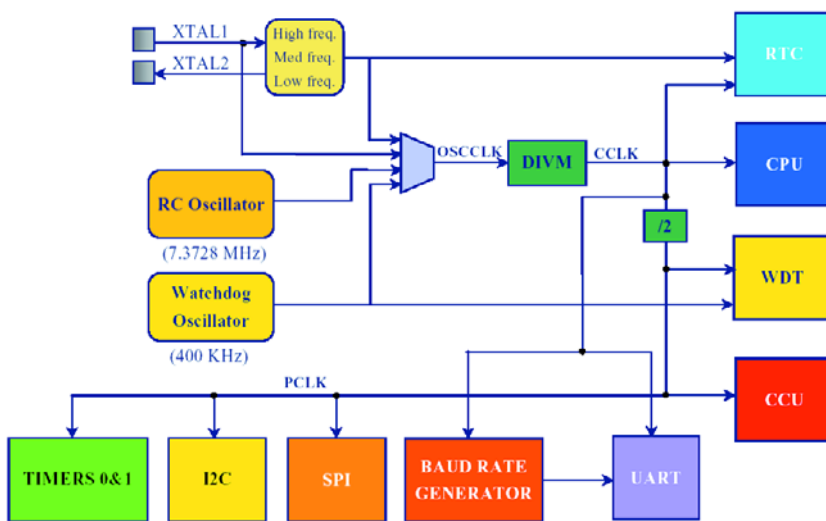


Figure 1. Complete overview of the clock distribution of the P89LPC900.

sureing the frequency at the clock output and fine-tune it with the aid of the TRIM register. Note that a nominal frequency of 7.3728 MHz was selected, which is a very appropriate value in order to generate all the known baud rates.

To be able to trade off performance and current consumption, the user can switch 'on the fly' to another frequency by changing the prescaler. Even the separate oscillator for the watchdog can serve as the clock source during moments

when performance is not that important, but current consumption plays a more significant role. **Figure 1** gives a complete overview of the clock circuitry.

The memory

The P89LPC900 is a complete embedded controller, which means that no external memory can be addressed (unless this is done via SPI or I²C, of course).

The data memory (RAM) of the

P89LPC900 partly consists of a section that can be **directly or indirectly** addressed (256 bytes) and, depending on the type, can be supplemented with a section of **external data**. The name may imply that this is external to the device, but in the case of the P89LPC900 it is definitely inside. The implications for the code are that the MOVX-instructions are required for access. A recent type, such as the P89LPC932, has, for example, 758 bytes, of which 256 bytes are direct/indirect and 512 are 'external'.

The code memory, based on flash technology, is not only repeatedly programmable, but can also be programmed **In System (ISP)** and even **In Application (IAP)**. For this purpose, Philips provides a bootloader which can be easily activated and communicates via the UART by default. An RS232 level shifter is all that is required to download the code from a PC. Esacademy provide the necessary software (Flash-Magic), which uses a simple user interface or command line (for production purposes, for example) to send the hex file to the processor.

The internal bootloader consists partly of built-in, low-level code that takes care of the interaction with the memory, and partly of a program in flash that deals with the communications aspect and the translation of the hex file. The programmer can quite easily use the low-level subset of the bootloader by writing to a few registers and executing a 'call' to a specific address. In this way you can write your own implementation of a bootloader, which, for example, does not communicate with the UART, but uses some other method chosen by the programmer. This new bootloader can replace the standard one. **Figure 2** shows how a call to the IAP subset is realised in practice.

The memory consists of pages of 64 bytes each. The pages assemble into 1k sectors. Pages and sectors can be individually erased, with an impressive erase time of only 2 ms. Also note that when programming the flash memory, no additional programming voltages are required. The second article will treat the subject of programming the flash memory in more depth. It may not be necessary to mention this, but the code memory can be programmed with a parallel programmer as well.

In addition to the usual data and code memories, the P89LPC900 also has an EEPROM memory (64 to 512 bytes). The EEPROM is accessed via a number of registers (pointer, data and status), so that it is not necessary to write specific communications code, as required for some controllers when the EEPROM is accessed via the I²C bus.

The P89LPC900 also has a number of special memory locations: the bootvector, the sta-

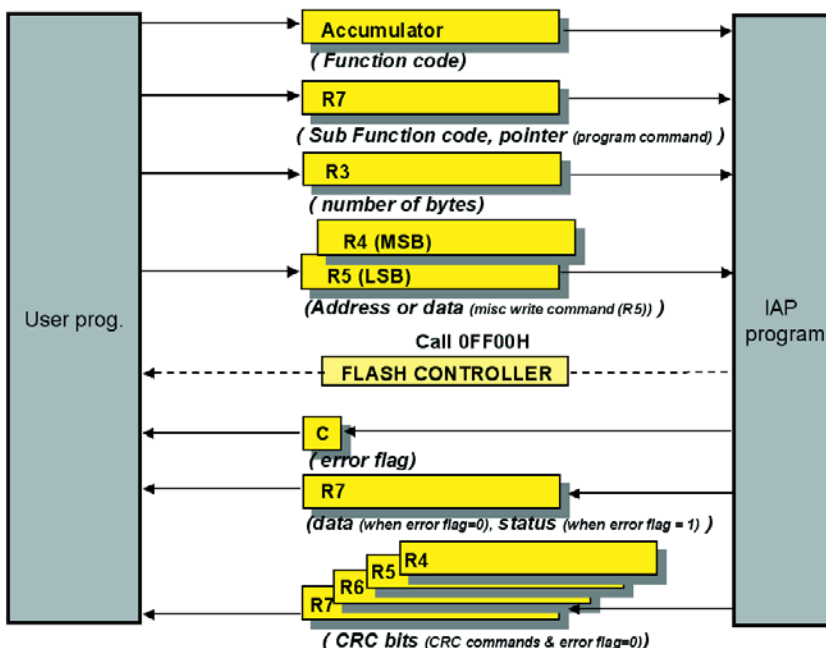


Figure 2. Here you can see how a 'call' has been realised to the IAP subset.

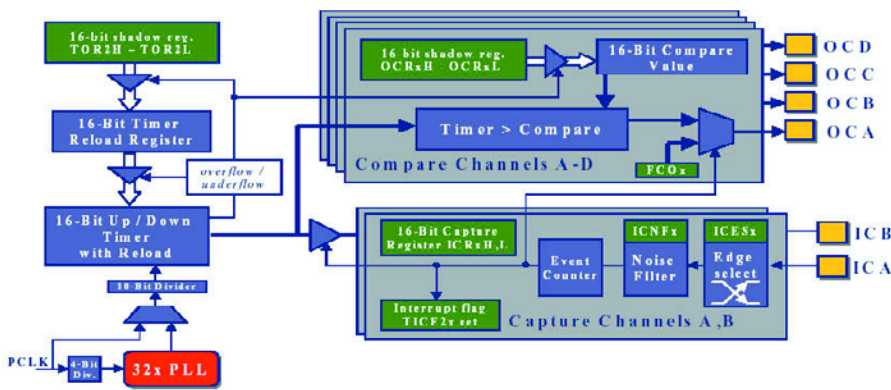


Figure 3. The user of the P89LPC932 can make use of 4 compare outputs and 2 capture inputs.

tus byte and the configuration byte. The bootvector determines the starting location when the chip is in boot mode. This vector is initially factory programmed to point to the internal bootloader. The status byte determines whether after reset the user code or the boot code is activated. In this case the initial value is again such that the boot code will be executed.

The last byte in this category is the configuration byte. It is used to select the desired **clock source**, the function of the **reset pin** and the system **supervisory settings**. The locations can also be accessed via ISP and IAP.

In- and outputs

Because the P89LPC900 has an internal **RC-oscillator** and a **Power-On-Reset** circuit, the basic rule is that the number of pins available for I/O is equal to the number pins on the package minus 2 for the power supply. In almost every case all the port pins can be programmed as **Input**, **Open-drain**, **Push-pull** or **Quasi-bidirectional**. Exceptions are the pins for the I²C bus and reset input. These have a different configuration because of their dual functionality. Once the processor has started, all pins are set to **input**. Two registers for each port (PxM1 and PxM2) together define the desired configuration. For the analogue functions there is a third register to ensure that the digital circuitry is completely decoupled from the relevant pin.

Naturally, a number of pins have more than one function (for communications, among others) and a few are suitable for interrupts. A group of eight port pins is available that can be used as a **keyboard-interrupt**. This mechanism is also provided with a **Pattern Match**,

which ensures that an interrupt (or a wake-up) is generated only when a specific pattern appears.

Five pins have, in addition to the digital I/O, an analogue comparator function. Two comparators with a selectable reference (internal or external) and switchable inputs can be used to compare signals and use the results to generate an interrupt. There are also versions with an 8- or 10-bit ADC.

When using the reset pin, take note of the alternative function of this pin. Even though the pin is defined as an input after power up, this pin is still sampled during reset. In the event that this pin is permanently 'low' the processor will not start up properly. In contrast to previous 80C51 controllers, the P89LPC900 has an active *low* reset!

System supervisory

The P89LPC900 is provided with all the features with regards to 'System Supervisory' functions. In the past a significant additional expense had to be made for external components that perform the same functions. The P89LPC900 has a **Power-On-Reset/Detect**, a **Low-Voltage Detect** and a **Watchdog**. The latter can, because of its separate oscillator, also function as an **Oscillator Fail Detect**. The activation of any of these functions does not necessarily result in a reset, but can generate an interrupt instead. A software-reset can then perform this task, if appropriate.

Power management

We mentioned before that the speed of the processor can be changed dynamically. For circuits that require even lower power consumption, there is the option of placing the processor in either **Idle** or **Power-down mode**. In both modes, the core itself is halted. The difference is that in Idle mode the peripherals continue to run, while in Power-down mode everything is stopped. The sources that can re-activate the processor differ for each mode. In Idle mode a timer can re-activate the processor, but in Power-down mode only an external/keypad or comparator interrupt can do this. Note that the watchdog can continue to run in Power-down mode, although this will increase the power consumption somewhat.

Timers and counters

The P89LPC900 has the well-known timer/counter 0 and timer/counter 1, which are more or less standard on every 80C51. New is the fact that the timers have an additional operating mode, which can be used as a 'poor man's PWM'. Registers THx and TLx together determine the frequency and duty cycle. There are, incidentally, versions (the P89LPC932 for example) that have a very comprehensive capture/compare unit with 16-bit resolution and a 32xPLL (refer **Figure 3**). This provides the user with 4 compare outputs, which permit all kinds of wave shapes, and two capture inputs. This module can be used to drive a stepper motor, for example, or control an H-bridge. A (noise filtered!) **Brake input** allows the option of hardware intervention in the event there is a need to immediately stop a stepper motor, for instance.

In addition to the timer/counters, the P89LPC900 is also provided with an RTC. This is a convenient way to keep track of elapsed time, without sacrificing another timer. **Figure 4** illustrates its operation.

Communications

The P89LPC900 has also acquired a range of new features in the area of communications. One of the more

notable new features is the presence of a separate **baud-rate generator** for the UART. In the past it was necessary to sacrifice a timer for this purpose, which limited further use of this timer to whatever frequency it had to run at.

The UART also has a break-detect mechanism. This makes it possible to, for example, activate the bootloader in a very simple manner. The user can now, without any effort, program the processor from within the programming environment. The Flash-Magic software has a facility to make this happen.

The I²C-interface complies with the 400-kHz standard and is byte-wide implemented, which greatly simplifies the code for the various I²C-modes and makes it less critical. A **state machine** can make use of the codes that are returned by the I²C hardware. This is an easy way to determine the progress of the I²C-transfer and decide what the next step should be. Philips provides a detailed description in the user manual as to which states belong with which mode (Master transmit/receive, Slave transmit/receive). By the way, the I²C-interface is compatible with earlier devices such as the P8xC554.

Even the SPI interface hasn't been forgotten. Communications at a speed of 3 Mbit is possible with a wide range of SPI devices. The SPI interface can be configured for either master or slave operation.

Interrupts

A large number of interrupts (15 on the P89LPC932) can be utilised to facilitate the smooth operation of the program. The interrupts have, by default, a pre-determined priority, but the user can divide them into four different levels. The interrupts can be activated or deactivated individually as well as globally.

Development tools

Because the P89LPC900 still has an 80C51 core, an assembler or compiler from the early days can still be used to write software for this family. The only thing that has changed is the required register file. This may be part of the assembler/compiler

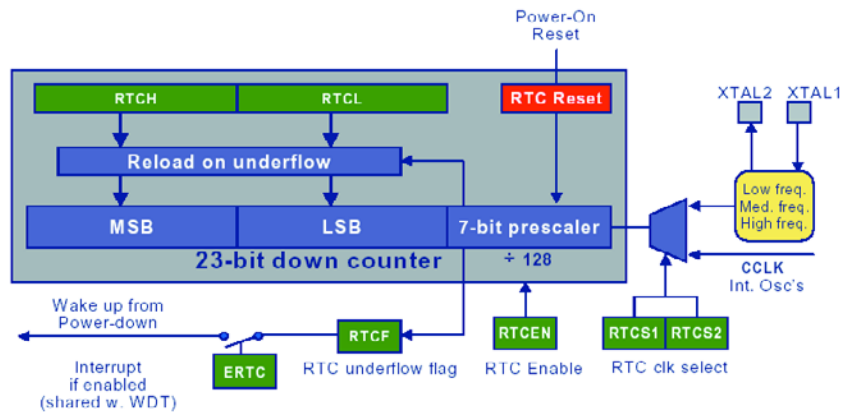


Figure 4. The standard RTC peripheral keeps track of elapsed time.

package or, alternatively, you can easily write one yourself.

On the Esacademy web site you should be able to find 'Code-Architect' by the time this issue of *Elektor Electronics* is published. This tool can generate online code for various peripherals, without having to consult the data sheet. The code can then be processed further in a C compiler.

Although the entire P89LPC900 family is available with flash memory and code is readily reprogrammed, an 'in-system-emulator' can drastically reduce the development time and generally make life much easier. There are a number of manufacturers that offer emulators for these controllers (Philips, ESS and Keil). The Keil emulator is remarkable. For the amount of only 199.00 euro (approx. £140), Keil supplies a fully-fledged emulator, including a C compiler with a 4 K limit. These tools all have a USB connection, so that, up to a point, an external power supply can be omitted. For hobbyists who do not want to spend this much money, the design to be described in Part 2 of this article will provide an alternative solution.

The family

The number of P89LPC900-derivatives is growing at a great pace. At the time of writing, it appears that Philips has chosen to fill the category of up to 28 pins first with the P89LPC900, (there are, after all, several flash-controllers in the higher category already) before producing the larger devices. Even controllers

with unusual peripherals such as CAN and USB have been planned.

Noteworthy are the 8-pin versions. Philips has decided to release a range of 8-pin devices into the market. It is simply the case that with 8 pins you don't get a great deal of choice regarding the peripherals. The various devices will therefore feature different combinations of peripherals.

Philips has done well to introduce a range of products on the market that have the same or greater functionality and are pin-compatible with the popular P87LPC76x, for example, which are only available in OTP versions.

The P89LPC900 are available in several different packages, but the emphasis will be on smaller packages such as the TSSOP (only 9.8 × 6.6 mm for the 28-pin version). The smaller technology (0.35 micron) and the fact that they are in-circuit programmable contribute to this. Fortunately, there are more practical package options for experimental work. In the project to be described in Part 2, the PLCC28 version has been used.

(030161-1)

Useful websites

www.philipsmcu.com

Philips 80C51-controllers and additional information such as application-notes and entry to the forum.

www.esacademy.com

Tools for the 80C51 such as Flash-Magic and Code-Architect.

www.altium.com

C-compiler, demo software limited to 2 K code can be downloaded.

www.keil.com

C-compiler, demo software limited to 2 K code can be downloaded.

www.pds51.com

Hardware tools for 80C51.

The content of this note is based on information received from manufacturers in the electrical and electronics industries or their representatives and does not imply practical experience by Elektor Electronics or its consultants.

TinyLogic ICs Type NC7SZ57/58

Configurable logic gates

Source: Fairchild Semiconductors

The thing about these TinyLogic integrated circuits from Fairchild's UHS series is that they allow up to five different gate combinations to be created, including versions with two identical inputs, or an inverting and a 'true' input.

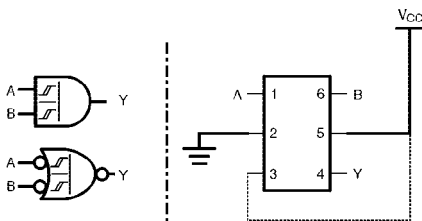


FIGURE 1. 2-Input AND Gate

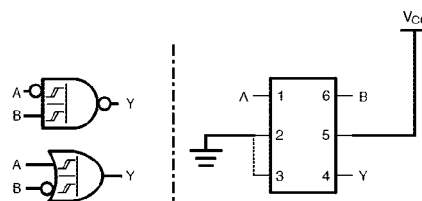


FIGURE 2. 2-Input NAND with inverted A Input

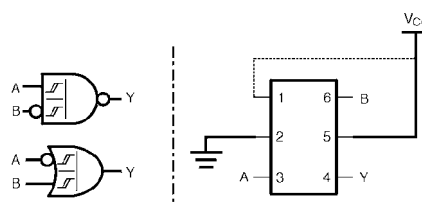


FIGURE 3. 2-Input NAND with inverted B Input

Figures 1-10. Overview of TinyLogic configuration options.

Fairchild's 'TinyLogic' family was originally introduced about five years ago. The series comprises a wide range of single and dual gate logic functions in a number of small SMD enclosures (hence 'tiny'). Compared with the commonly seen 'small outline' cases of similar CMOS ICs, TinyLogic devices offer considerable space savings on printed circuit boards.

All members of the TinyLogic family are marked by high speed and low power consumption. These features, in combination with their small size, make the devices eminently suitable for application in portable equipment like portable CD-players, GSM phones, PDAs and laptop computers, where space is at a premium.

Within the family, a number of sub-series may be identified: HS, HT, UHS and, recently, the ULP series. The respective acronyms stand for 'high speed', 'high-speed TTL compatible', 'ultra-high speed' and 'ultra-low power'. The devices discussed in this article belong in the UHS series which is marked by a supply voltage range of 1.65 V to 5.5 V and

very high switching speeds. Compared with 'ordinary' HC logic the speed of the NC7SZ devices is more than twice as high at just 2.8 ns typical (at 5 V supply voltage). Another extremely welcome characteristic of these ICs is that their outputs can supply a current of up to 32 mA. This makes them suitable for direct driving of LEDs, optocouplers and low-power relays.

One IC, five functions

Notwithstanding the great value of the above mentioned specifications, undoubtedly the most striking thing about the NC7SZ57 and -58 is that each of these can be configured for five different logic functions. Simply by 'rearranging' the connection of the output pins, the IC allows all standard logic functions with two inputs to be implemented. The inset table and the associated **Figures 1 through 10** illustrate how this may be achieved in practice.

Figures 1 through 5 cover the NC7SZ57 and Figures 6 through 10, the NC7SZ58. All inputs feature hysteresis.

Table I. Configurations

2-Input Logic Function	Type	Circuit
2-Input AND	NC7SZ57	Figure 1
2-Input AND with one inverting input	NC7SZ58	Figure 7, 8
2-Input AND with two inverting inputs	NC7SZ57	Figure 4
2-Input NAND	NC7SZ58	Figure 6
2-Input NAND with one inverting input	NC7SZ57	Figure 2, 3
2-Input NAND with two inverting inputs	NC7SZ58	Figure 9
2-Input OR	NC7SZ58	Figure 9
2-Input OR with one inverting input	NC7SZ57	Figure 2, 3
2-Input OR with two inverting inputs	NC7SZ58	Figure 6
2-Input NOR	NC7SZ57	Figure 4
2-Input NOR with one inverting input	NC7SZ58	Figure 7, 8
2-Input NOR with two inverting inputs	NC7SZ57	Figure 1
2-Input XOR	NC7SZ58	Figure 10
2-Input XNOR	NC7SZ57	Figure 5

User friendly

The TinyLogic ICs come in an 'SC70' SMD case which is only 2 mm long! The pin connections are shown in **Figure 11**, while the pin states are listed in a separate table.

Remarkably, and in contrast with what we would expect, each TinyLogic gate has its own case. It is also unusual that the case does not have a mark to identify pin 1. Here,

the print on the device tells us the location of pin 1. If the print is legible, pin 1 is below at the left.

When designing a PCB and constructing a circuit, there are certain advantages to having just one gate in one integrated circuit. For example, the logic gate may be inserted directly in the signal line, without having to run tracks up and down the board to a central IC containing, say, four such gates. The result is

Table 2. Electrical Data

Absolute Maximum Ratings

DC Electrical Characteristics	
Supply Voltage (V_{CC})	-0.5 V to +7.0 V
DC Input Voltage (V_{IN})	-0.5 V to +7.0 V
DC Output Voltage (V_{OUT})	-0.5 V to +7.0 V
DC Input Diode Current (I_{IK}) @ $V_{IN} \leq 0.5$ V	-50 mA
DC Output Diode Current (I_{OK}) @ $V_{IN} \leq -0.5$ V	-50 mA
DC Output Source/Sink Current (I_{OUT})	± 50 mA
DC VCC or Ground Current (I_{CC} / I_{GND})	± 50 mA
Storage Temperature Range (T_{STG})	-65°C to +150°C
Max Junction Temperature under Bias (T_J)	150°C
Lead Temperature (T_L) (Soldering, 10 seconds)	260°C
Power Dissipation (P_D) @ +85°C SC70-6	180 mW

Recommended Operating Conditions

Supply Voltage Operating (V_{CC})	1.65 V to 5.5 V
Supply Voltage Data Retention (V_{CC})	1.5 V to 5.5 V
Input Voltage (V_{IN})	0 V to 5.5 V
Output Voltage (V_{OUT})	0 V to V_{CC}
Operating Temperature (T_A)	-40°C to +85°C
Thermal Resistance (θ_{JA}) SC70-6	350°C/W

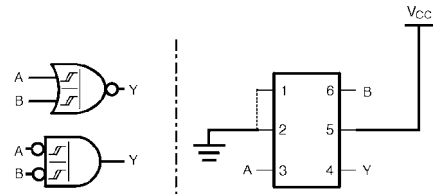


FIGURE 4. 2-Input NOR Gate

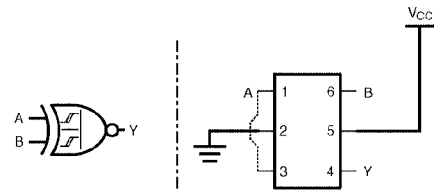


FIGURE 5. 2-Input XNOR Gate

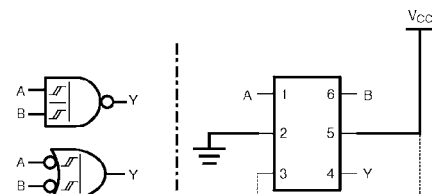


FIGURE 6. 2-Input NAND Gate

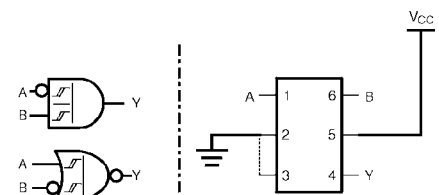


FIGURE 7. 2-Input AND with inverted A Input

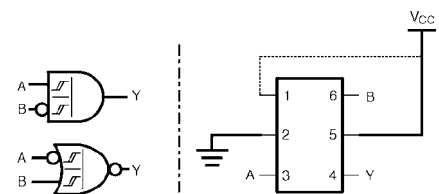


FIGURE 8. 2-Input AND with inverted B Input

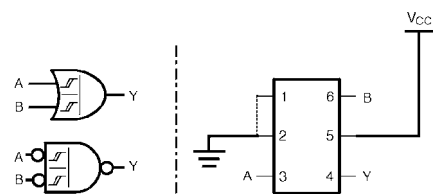


FIGURE 9. 2-Input OR Gate

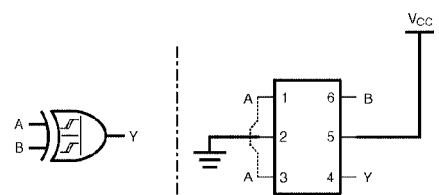


FIGURE 10. 2-Input XOR Gate

shorter connections with all the well-known benefits for the actual operation of the circuit. Besides, PCB designers have more freedom in doing the board layout, which does not require a total redesign following every minor change to the circuit.

Conclusion

Summarizing, we can say that these TinyLogic gates are a useful addition to the usual collection of logic components. TinyLogic devices are small, fast, economical and their ease of configuration makes them really versatile.

To close off this short article, we should not forget to mention the main electrical specifications of the NC7SZ57 and -58; see the second table.

Those of you with a deeper interest in these unusual but useful ICs are referred to the TinyLogic datasheets published by Fairchild. These may be downloaded free of charge from this url:

www.fairchildsemi.com

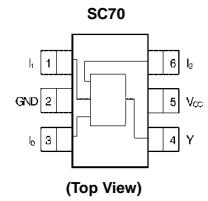
Pin Descriptions

Pin Name	Description
I_0, I_1, I_2	Data Inputs
Y	Output

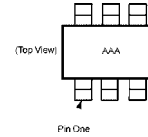
Function Table

Inputs			NC7SZ57				NC7SZ58					
I_2	I_1	I_0	Y	(I_0)	(I_2)	(I_1)	(I_2)	Y	(I_0)	(I_2)	(I_1)	(I_2)
L	L	L		H				L				
L	L	H		L				H				
L	H	L		H				L				
L	H	H		L				H				
H	L	L		L				H				
H	L	H		L				H				
H	H	L		H				L				
H	H	H		H				L				

H HIGH Logic Level L LOW Logic Level

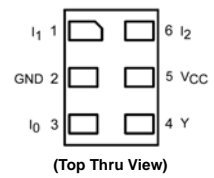


(Top View)



AAA Product Code Top Mark - see ordering code
Note: Orientation of Top Mark determines Pin One location. Read the top product code mark left to right, Pin One is the lower left pin (see diagram).

Pad Assignment for MicroPak



(Top Thru View)

020353 - 12

(020353-1)

Figure 11. Case, pinout and pin states.

Precision Measurement Central (3)

Part 3: Flash for all purposes

By J. Wickenhäuser

www.wickenhaeuser.com

It would appear that extensive study of the datasheets is required to fully exploit the galaxy of possibilities offered by the MSC1210 family of microcontrollers. Not so for Elektor readers! In this article you will find everything you need to know to be able to make the best possible use of the 32-kB Flash memory contained in the MSC121x.

At first glance the members of the MSC121x family appear to have the usual 8051 memory architecture: up to 64 kBytes of code space (program memory) and the same amount for data (RAM) may be addressed by the core. In accordance with the so called Harvard architecture, there are memory areas for code and data (Figure 1). On closer inspection we see that various types of overlap are possible with the MSC121x. Even if rarely required, these options allow the internal 1-kB RAM to be mapped into the program as well as the data memory. In this article we aim to show how the Flash memory area may be tailored to your exact requirements.

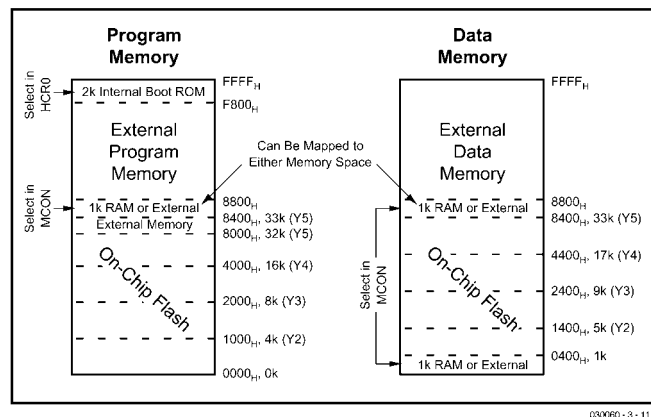


Figure 1. Memory architecture.

Flash partitioning

The above operation is called “Flash partitioning” by Texas Instruments. Essentially, it means that during the code download you are able to decide how much Flash memory you want to allocate to the program and how much to the data. Depending on the exact controller, up to seven options are available to choose from (Figure 2).

In practice, the division has to be set up as you write the source code, that is, at an early stage in the program development. As far as we are aware there are hardly any other processors on the market that support this feature. The allocation to either program or data memory ensure that access to wrong addresses in the data range (like an inadvertent ‘erase’ operation) can not affect the exe-

cutable program, which is obviously a bonus when talking about program security and reliability. Moreover, the data blocks are at fixed addresses (Figure 3), which may simplify their management in software.

As a matter of course, write access to the code area may be allowed, as in, for example, systems that require software to be conveyed ‘in application’ (IAP). A practical example is the Precision Measurement Central (Elektor Meter) where the whole download operation requires no more than a few components, two jumpers and a ‘fully loaded’ RS232 interface. Just imagine, your equipment needs

new firmware and there is no other connection to it than RS485, Ethernet or infrared — who will pull any jumpers? It is exactly in such cases that IAP is the perfect option.

The Flash memory comprises up to 256 sectors of 128 bytes each. According to the datasheets, it is capable of sustaining up to a million write or erase operations per memory cell (the number or read operations is indefinite). No further actions are required to be able to read the Flash memory. The code always starts at address \$0000 and data at \$0400 (according to Figure 3, the additional 1 kB RAM is located

HCR0	MSC1210Y2		MSC1210Y3		MSC1210Y4		MSC1210Y5	
DFSEL	PM	DM	PM	DM	PM	DM	PM	DM
000	0kB	4kB	0kB	8kB	—	—	—	—
001	0kB	4kB	0kB	8kB	—	—	0kB	32kB
010	0kB	4kB	0kB	8kB	0kB	16kB	16kB	16kB
011	0kB	4kB	0kB	8kB	8kB	8kB	24kB	8kB
100	0kB	4kB	4kB	4kB	12kB	4kB	28kB	4kB
101	2kB	2kB	6kB	2kB	14kB	2kB	30kB	2kB
110	3kB	1kB	7kB	1kB	15kB	1kB	31kB	1kB
111 (default)	4kB	0kB	8kB	0kB	16kB	0kB	32kB	0kB

NOTE: When a 0kB program memory configuration is selected program execution is external. "—" is reserved.

030060 - 3 - 12

Figure 2. Possible size of the Flash partitions.

HCR0	MSC1210Y2		MSC1210Y3		MSC1210Y4		MSC1210Y5	
DFSEL	PM	DM	PM	DM	PM	DM	PM	DM
000 (reserved)	—	—	—	—	—	—	—	—
001	0000	0400-13FF	0000	0400-23FF	0000	0400-43FF	0000	0400-83FF
010	0000	0400-13FF	0000	0400-23FF	0000	0400-43FF	0000-3FFF	0400-43FF
011	0000	0400-13FF	0000	0400-23FF	0000-1FFF	0400-23FF	0000-5FFF	0400-23FF
100	0000	0400-13FF	0000-0FFF	0400-13FF	0000-2FFF	0400-13FF	0000-6FFF	0400-13FF
101	0000-07FF	0400-0BFF	0000-17FF	0400-0BFF	0000-37FF	0400-0BFF	0000-77FF	0400-0BFF
110	0000-0BFF	0400-07FF	0000-1BFF	0400-07FF	0000-3BFF	0400-07FF	0000-7BFF	0400-07FF
111 (default)	0000-0FFF	0000	0000-1FFF	0000	0000-3FFF	0000	0000-7FFF	0000

NOTE: Program memory accesses above the highest listed address will access external program memory.

030060 - 3 - 13

Figure 3. Ranges of the Flash partitions.

between \$0000 and \$03FF of the data memory area).

Writing to the Flash memory

For Flash writing we need to open our bag of tricks. The *2k Internal*

Boot ROM shown in Figure 1 contains a number of useful routines including one that gives software access to the Flash memory. A header file called ROM1210.H that comes with the uC/51 Compiler is the key to this access.

From a technical point of view, it

is essential to understand the difference between the two memory types Flash and E2PROM. In an E2PROM each memory cell is individually addressable, while in a Flash memory only complete sectors can be erased, whereupon all addresses will read \$FF (i.e., all bits are at 1), and writing is limited to 0's. One of the original Texas Instruments demo programs simulates an E2PROM by first copying the relevant sector into RAM, clearing the sector, inserting the new value into RAM and then copy the complete RAM back into the cleared sector again.

The HCR0 register serves to allocate the Flash part. This register is only be written to during the 'official' download, when it is mapped to address \$807 in the program code area. There are currently two configuration registers, HCR0 and HCR1, but the latter is not required for the above operation.

The source code files go to demonstrate the simplicity of using the uC/51 Compiler to employ hardware-like settings like the HCR registers or the reserving of memory in the code range (from \$100).

The value \$9E for HCR0 reserves 1 Kbyte Flash for data and allows unlimited access to the Code and Data Flash. It is essential to set the time registers of the MSC121x to the right values (depending on the quartz crystal frequency). If the baud rate is also right at that point, things can start happening!

As before, use the Umshell utility to load the program onto the MC1210 board using **run** (Figure 4). Each character you type is copied into the two memories which are subsequently displayed (different brackets!). Note however that if you exceed 40 characters you run the risk of corrupting your program, because the reserved range comprises only 40 bytes.

The full source code files are found in the uC/51 Compiler from version V1.10.10 onwards at the indicated Internet address. From this version onwards, HEX files are additionally compressed for the Flash download.

(030060-3)

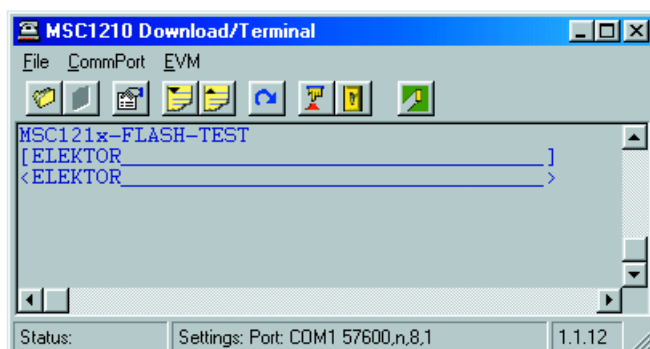


Figure 4. Flash Test by way of UmShell.

Internet links

<http://www.wickenhaeuser.de>

Here you will find the latest version of the uC/51 Compiler, drivers and demos.

www.lmphotronics.com/forum/forumdisplay.php?fid=68

A new forum for users of the uC/51 Compiler.

SuperCap Battery

Power from a GoldCap

By B. Kainka

GoldCap capacitors offer an interesting alternative power source when compared to conventional disposable or even rechargeable batteries. They can be charged very rapidly and can also deliver a high peak output current. Their voltage rating however is quite low so a little electronic assistance is necessary to raise the output voltage to a more useful level.

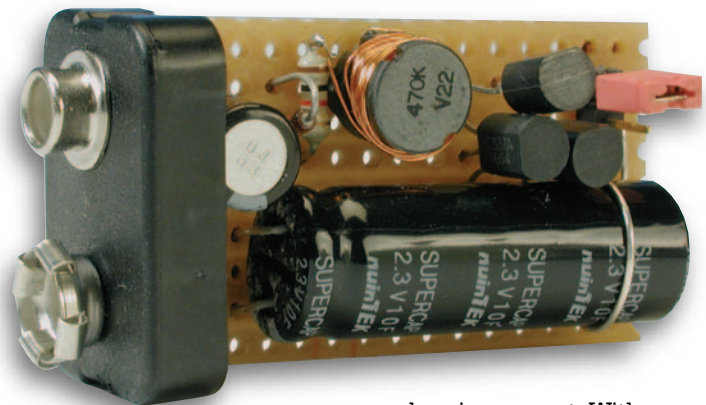
PP3 (6F22) type 9 V batteries are often used in small portable equipment that require very little current and may only be used intermittently. Under these conditions its often the case that the battery is flat just when you urgently need to use the equipment. NiCd rechargeable cells are not a good choice in these applications because their self-discharge characteristics are much worse than dry cells and often there is no charge left after a long time in storage. Capacitors offer an interesting alternative power storage device, they can be charged very quickly and they retain their charge for years. SuperCaps are small outline capacitors offering huge storage capacity measured in farads rather than the microfarads (μF) that we are used to. The maximum supply voltage of 2.3 V means that some form of voltage multiplier is necessary to increase voltage up to a more useful 9 V. Using a 10 F, 2.3 V SuperCap specified here, the complete circuit can be built into a package the same size as a PP3 cell. Compared to rechargeable cells a capacitor can handle a very high peak current for both charging and discharging and has a very low self-discharge characteristic (it keeps its charge). A capacitor is an almost ideal energy storage medium as can witnessed for example by its universal use in camera flash equipment. There are no batteries (dry or rechargeable) of the equivalent size that can deliver so much energy in such short a time. Also there are very little loses involved in the charge/discharge cycle. Now the bad news; unlike a battery a capacitor has a very simple formula relating its voltage to charge stored so that when charge is removed, output voltage falls. In this respect a rechargeable cell is

much more user-friendly, it keeps the output voltage relatively constant over the discharge cycle until the charge is used up when the output voltage falls more sharply. A capacitor will therefore need some form of voltage regulation circuit before it can act more like a battery.

The switching regulator

The circuit shown here in **Figure 1** uses a small outline SuperCap with a capacitance of 'just' 10 F (yes that's 10,000,000 μF !) with a maximum operating voltage of 2.3 V. A switching voltage regulator is used to pump up the capacitor voltage to 9 V. A small switch or jumper is fitted to the circuit to ensure that the capacitor retains its charge over time. An LM317T voltage regulator (IC1) is used on the input to ensure that the capacitor can be recharged from a wide range of voltage sources whilst protecting the capacitor from over-voltage. This IC has built-in over-current and over-temperature protection. The circuit can be charged from a standard mains adapter where the internal resistance of the adapter will limit the

charging current. With an input current of 1 A the capacitor will be fully charged in 20 s! The rest of the circuit is a switching regulator containing a small ferrite coil or choke (L1). The 470 μH choke consists of approximately 20 turns of enamelled copper wire wound around a ferrite core. The resistance of the winding should be less than 1 Ω . The two ends of the winding are soldered to connecting pins in the body of the choke. **Figure 2** shows the second winding of 20 turns that needs to be wound over the top of the existing winding to provide feedback for the oscillator (wind in the same direction as the first winding). Alternatively the existing winding can be removed from the choke and replaced by a 40 turn centre-tapped winding of enamel coated copper wire (0.5 mm diameter or 24/26 SWG is suitable). The choke has now been converted into a small transformer. Oscillator operation has proved to be reliable and runs with an input voltage as low as 0.5 V.



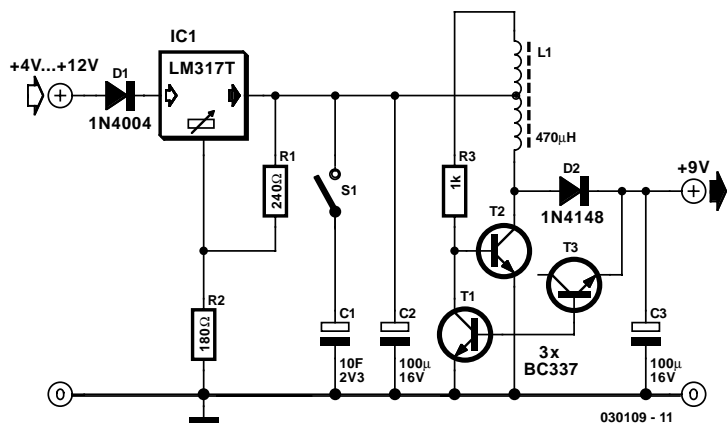


Figure 1. Circuit diagram of the Goldcap rechargeable.

The output voltage is regulated to provide approximately 9 V. Transistor T3 uses its reverse biased base-emitter junction to act as a zener diode, providing a voltage reference to regulate the output. The zener voltage of a small signal NPN transistor is around 8 V in this configuration (an 8 V zener diode could be substituted for T3). The circuit produces an output of 8.5 V with an input voltage of 2.3 V while at 0.7 V the output falls to 8.4 V.

Two of the most important criteria for any switching regulator are its output current capability and its efficiency. A 1 kΩ load resistor was used for testing. The circuit took 50 mA with an input voltage of 2.3 V giving a power consumption figure of 115 mW. The output voltage of 8.5 V is the same as the output voltage with no load. Current in the load is 8.5 mA giving an output power of 72.3 mW and this equates to an efficiency of 63 %.

Output current is about 1 mA with a load resistance of 8.2 kΩ. The output voltage remains constant until the SuperCap voltage falls to 1 V by

which time the capacitor will have given up 80 % of its stored energy. As the voltage continues to fall the output also sinks until at 0.6 V the output voltage is 4.8 V.

As with any switching regulator, the input current increases as the input voltage falls. This type of regulator is more efficient than a standard linear regulator so that smaller loads result in lower power consumption. Under no-load conditions the converter consumes just 2 mA with an input voltage of 2.3 V. Lightly loaded, the circuit will run for up to three hours before it needs recharging.

The SuperCap used in this circuit is produced by the Korean company NuinTEK and has a package outline 10 mm diameter by 30 mm. Farnell stock a similar SuperCap produced by Panasonic, this device is a little larger with a body diameter of 18 mm and will not unfortunately fit into a PP3 outline. Alternatively there are 10 F GoldCaps rated at 2.5 V, these are more generally available but the case size is again bigger than the NuinTEK capacitor.

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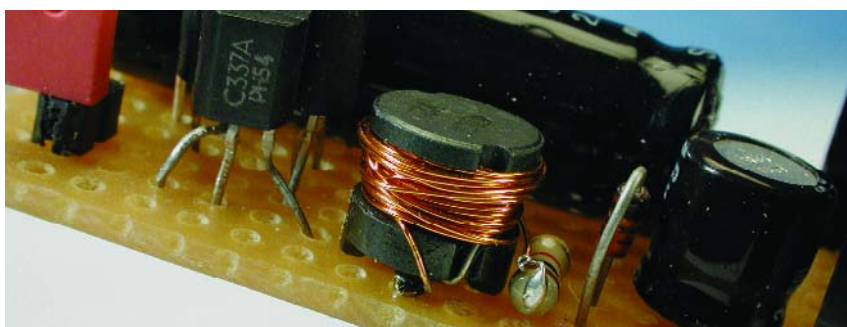


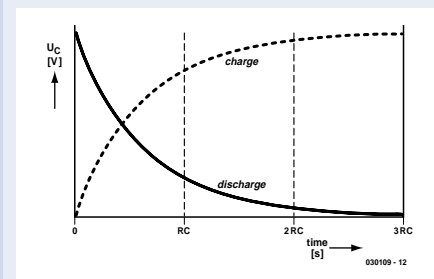
Figure 2. Coil L1 winding details.

Capacitance measurement

Most capacitance meters are not capable of measuring capacitors of 1 F and more, so what can you do when you need to measure a SuperCap or GoldCap capacitor? Well the answer is not that difficult, all you need is a timer (a watch will do), a voltmeter and a load resistor.

The voltage stored on a capacitor does not fall linearly when it is connected to a constant resistance but exponentially. The curve gets flatter with time because as the voltage falls so the current in the load gets smaller causing less and less charge to be taken from the capacitor. Theoretically it will never be totally discharged, with time the discharge curve can approach but never actually achieve zero volts. This simple capacitor/resistor circuit is said to have a time constant $\tau = R \cdot C$. Where τ is seconds, R ohms and C farads. After each time constant the output voltage falls by 37% of its starting value and in practice falls to within 1% of its final value after 5 time constants. A 10 F SuperCap together with a 100 Ω resistor has a time constant $\tau = 100 \Omega \times 10 F = 1000 s$ (16 Minutes and 40 Seconds). Armed with this information we can now start measuring!

Starting with the capacitor charged up to 2 V we can work out the value of capacitance by measuring the time it takes for the voltage to fall by 37% (37% of 2 V equals 0.74 V so the voltage across the capacitor will be 1.26 V after one time constant) now plugging this time into the formula $C = \tau/R$ will give us the value of C. In our test the voltage took 1200 s to fall to 1.26 V so the capacitance measured is 12 F.



Don't be tempted to use too low a value of load resistance to speed up the test; you could end up burning your fingers (literally). The SuperCap can deliver a peak current of around 6 A, this equates to a lot of power (heat) dissipated in low resistance loads. The capacitor is also not a perfect device and like all capacitors has a certain amount of internal series resistance. The smaller the value of load resistance the greater will be the effect of dissipation in this internal resistance.

DC-Operated Stepper Motors

With unidirectional rotation and slow speed

By A. Schilp

This circuit has been designed specifically to allow a small stepper motor to rotate in one direction at a very slow speed. The main application is in the area of model making.

A circuit like the one described here can be very useful when little power and slow speeds are required, but a gearbox is undesirable. Some applications to consider are a radar antenna on a model ship, which rotates at about 30 times per minute, or the vanes of a windmill as part of a model railway, which turns at no more than about 20 revolutions per minute. The driver circuit described here provides a speed from about 5 to 100 rpm and is therefore ideal for these types of applications.



(Photo: courtesy Kenteq)

Step size

In practice, stepper motors are not usually intended for rotating applications, but are used for positioning. A stepper motor doesn't actually rotate in the usual sense; it makes small steps instead. You have to take into account the specification for power supply of the windings and the number of degrees per step. When the step angle is greater than 1.8 degrees, that is, less than 200 steps per revolution, the individual steps can almost always be discerned and a smooth, slow turning motion cannot be realised with this circuit. The best results will be obtained with a stepper motor with a step angle of 0.9 degrees, that is, 400 steps per revolution. Higher steps per revolution are rare in practice.

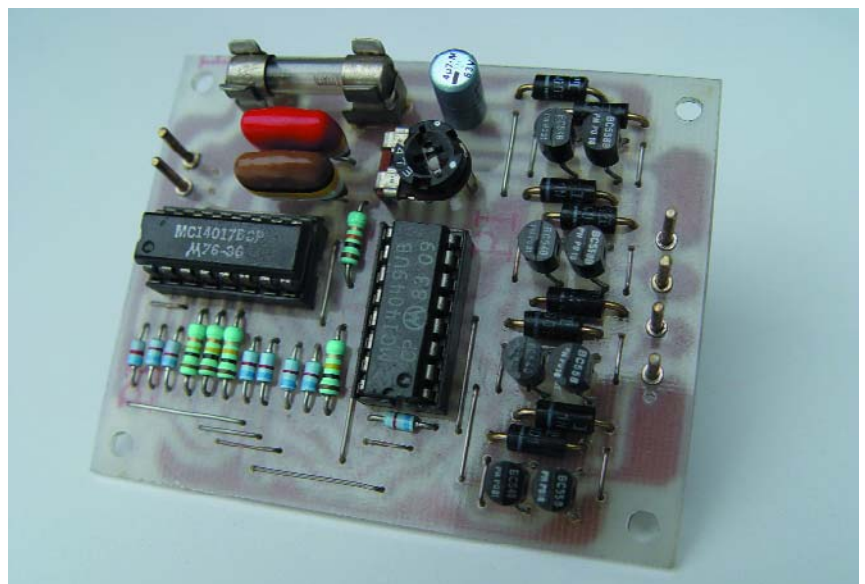
Schematic

As can be seen from the schematic shown in **Figure 1**, the complete driver circuit consists of little more than two ICs and eight transistors. There are no 'difficult' parts, they are all, with-

out exception, standard components.

The rotational speed of the stepper motor is determined by the posi-

tion of P1, which adjusts the frequency of the clock pulse generator consisting of IC1d, IC1e, R5, P1 and



C3. The value of C3 may be changed to suit, but above 100 revolutions per minute, most stepping motors will refuse to cooperate. There is no minimum speed.

The clock pulse is routed to the clock input (pin 13) of IC2. This is a decade counter of the type 4017, which has been configured so that at each clock pulse, one of the outputs Q0 (pin 3) through to Q7 (pin 6) goes high in succession. Once Q7 has been 'high', the counter resets itself and the cycle repeats. The truth table shown in **Figure 2** illustrates the operation.

R1 through to R4 and D9 through to D16 form a matrix that converts the eight sequential states of the counter to driver signals for the winding terminals A, B, C and D. The truth table lists the voltage potentials at the winding terminals, as well as the resulting coil voltages U_{AB} and U_{CD} .

The drivers consist of T1 through T8, D1 through D8 and IC1a to IC1d at the coil ends A, B, C and D. The ends of the windings are connected to either power or ground, depending on the logic level at the input of the inverter. It is self-evident that the voltage across the winding is equal to zero when both ends are connected to the same power supply rail.

Remarks

Many stepper motors have windings with a centre tap. This connection must be left open because it is not used with this bipolar drive circuit. Implementations without centre taps can be recognised from the fact that they only have four connecting wires. Designs with centre tap have five or six wires. Note that motors with five wires are not suitable for this application, because the centre taps have been connected internally.

The circuit described here is intended for the smallest possible stepper motor and the lowest possible supply voltage. The larger the motor and the higher the power supply voltage, the more work the driver transistors T1 through T8 have to do. Fuse F1 limits the current through the transistors and diodes. If the drivers have to switch higher currents, then it is desirable to replace T1, T3, T5 and T7 with the Darlington transistor BD677 and use its complement, the BD678, to replace T2, T4, T6 and T8. These transistors possess integrated avalanche diodes, so diodes D1 through D8 may be omitted.

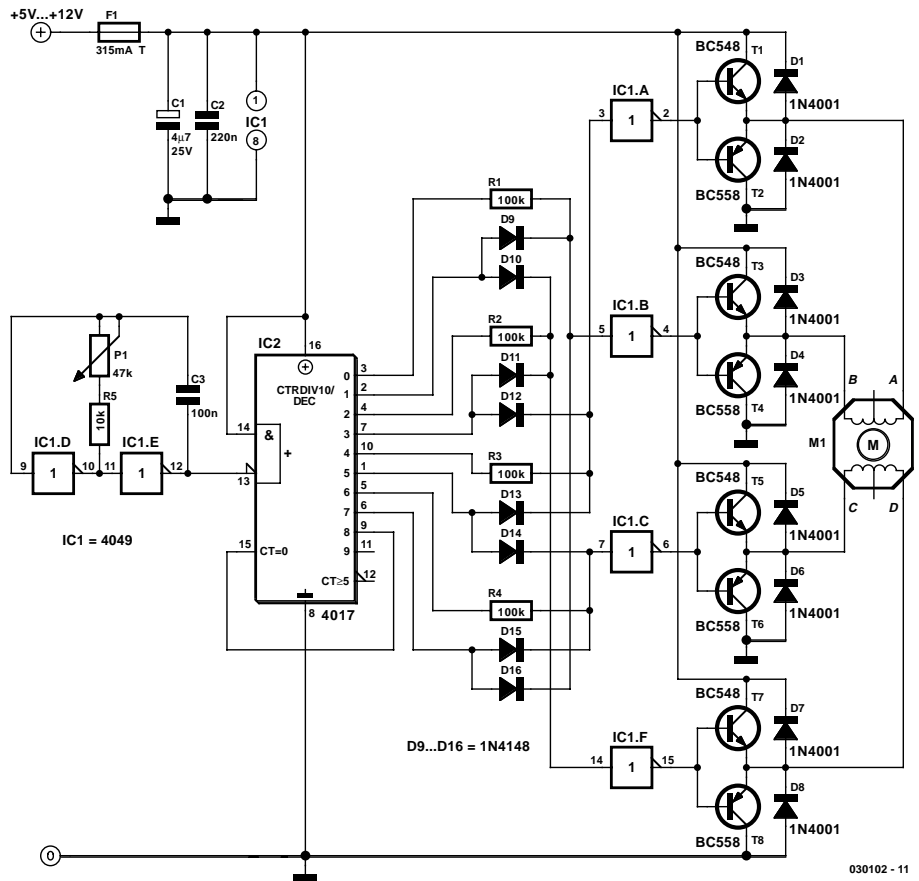


Figure 1. The driver circuit comprises very few parts.

CLK IC2 ø 13	Q0 IC2 ø 3	Q1 IC2 ø 2	Q2 IC2 ø 4	Q3 IC2 ø 7	Q4 IC2 ø 10	Q5 IC2 ø 1	Q6 IC2 ø 5	Q7 IC2 ø 6	A	B	C	D	U_{AB}	U_{CD}
0	1	0	0	0	0	0	0	0	+V	GND	+V	+V	↑	
1	0	1	0	0	0	0	0	0	+V	GND	+V	GND	↑	↓
2	0	0	1	0	0	0	0	0	+V	+V	+V	GND		↓
3	0	0	0	1	0	0	0	0	GND	+V	+V	GND	↓	↓
4	0	0	0	0	1	0	0	0	GND	+V	+V	+V	↓	
5	0	0	0	0	0	1	0	0	GND	+V	GND	+V	↓	↑
6	0	0	0	0	0	0	1	0	+V	+V	GND	+V		↑
7	0	0	0	0	0	0	0	1	+V	GND	GND	+V	↑	↑

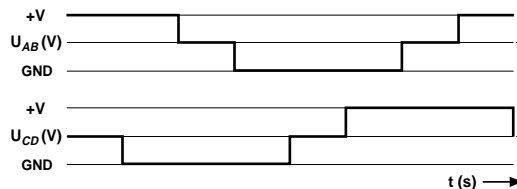


Figure 2. This truth table illustrates the operation of the stepper motor driver.

The author has designed a simple PCB for the circuit. The layout is unfortunately not available, but considering the simplicity, the driver circuit can also be easily built on a piece of prototyping board.

R-C Network Measurement with BASCOM AVR

Better accuracy from a new command

By J. C. Feltes

The popular BASIC compiler for AVR microcontrollers has been expanded with a very useful command. Instead of using the inaccurate GetRC0 command, you're now able to establish R-C values by means of two port pins.

When a microcontroller has no analogue/digital converter, you need to resort to tricks if you want to measure the value of a resistor or a capacitor in an R-C network. For this purpose, BASCOM AVR has the command GetRC available. With reference to **Figure 1**, the R-C network connected to port line D4 should have a 100-nF capacitor while the value of the resistor to be measured is between 0 and 10 kΩ.

GetRC arranges for the capacitor to be charged in one go, and discharged in small steps until the Low state is reached. The number steps are counted and stored in a 'word' variable (**Listing 1**).

Of course, the display should be prepared and configured to enable the micro to show the measured value. Unfortunately, sound as the above theoretical concept may appear, it fails miserably in practice. For example, two successive measurements (10 kΩ and 4.7 kΩ) using the fixed 100-nF capacitor produced the following values on the display: 756 and 508 respectively.

In order to improve the accuracy an experiment could be set up in which the GetRC command is mimicked by a series of discrete

steps. A suggested software tool is shown in **listing 2**. In the first part, the capacitor is charged during 50 ms to make sure it is 'full'. Next, the relevant port pin is declared an 'output' and pulled Low. Each of the commands takes 0.5 μs to complete (assuming a 4-MHz clock used on an AT90S8535 which requires two machine cycles per instruction). Next, the port pin is configured as an input again and the software runs a check to see if the capacitor voltage has dropped Low already. Switching from output to input again takes 0.5 μs. If the capacitor is not yet discharged, the variable is incremented (0.4 μs) before the return occurs.

Listing 2 shows different resistor values entered for GetRC and \$\$GetRC. Clearly, the measurement principle only yields reasonably reliable results for resistor values up to about 10 kΩ. At higher resistance values, the measured values even seem to drop (probably because of various leakage currents). However,

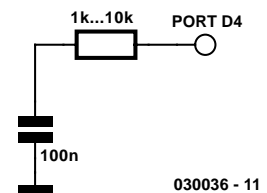


Figure 1. Circuit configuration for the original GetRC command and \$\$getrc.bas.

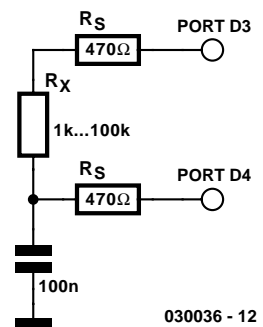


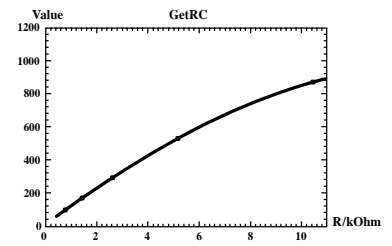
Figure 2. Two port pins are needed for a precision measurement using \$\$getrc2.bas.

Listing 1: measurement using original command GetRC

```

Config Lcdpin = Pin , Db4 = Portc.4 , Db5 = Portc.5 , Db6 = Portc.b ,
                Db7 = Portc.7 , Rs = Portd.7 , E = Portd.6
Config Lcd = 20 * 2
Dim W As Word
Do
    W = Getrc(pind , 4)
    Cls
    Lcd W
Loop
    
```

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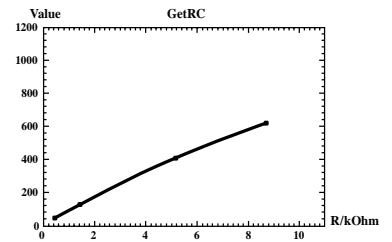
Listing 2: measurement using \$\$getrc.bas according to MSC method

```

Config Lcdpin = Pin , Db4 = Portc.4 , Db5 = Portc.5 , Db6 = Portc.b ,
                Db7 = Portc.7 , Rs = Portd.7 , E = Portd.6
Config Lcd = 20 * 2
Dim W As Word

Do
    Config Pind.4 = Output
    Portd.4 = 1
    Waitms 50
    W = 0
    Do
        Config Pind.4 = Output
        Portd.4 = 0
        Config Pind.4 = Input
        If Pind.4 = 0 Then Exit Do
    Else
        W = W + 1
        If W > 65000 Then Exit Do
    End If
    Loop
    Cls
    Lcd W
Loop
    
```

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Listing 3: measurement using \$\$getrc2.bas for 2 port pins

```

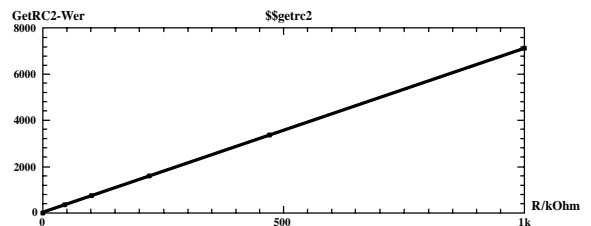
Declare Sub Getrc2(w As Word)
Config Lcdpin = Pin , Db4 = Portc.4 , Db5 = Portc.5 , Db6 = Portc.b ,
                Db7 = Portc.7 , Rs = Portd.7 , E = Portd.6
Config Lcd = 20 * 2
Dim W As Word

Do
    Call Getrc2(w)
    Cls
    Lcd W
Loop

Sub Getrc2(w)
    Config Pind.4 = Output
    Config Pind.3 = Output

    Portd.4 = 1
    Portd.3 = 1
    Waitms 10
    W = 0
    Config Pind.4 = Input
    Portd.4 = 0
    Portd.3 = 0
    Do
        If Pind.4 = 0 Then Exit Do
    Else
        W = W + 1
        If W > 65000 Then Exit Do
    End If
    Loop
    Config Pind.4 = Output
    Portd.4 = 1
    Portd.3 = 1
End Sub
    
```

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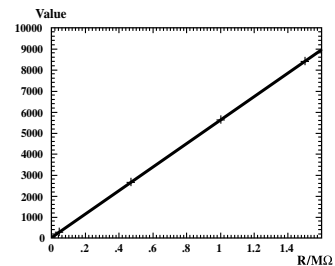
Listing 4: measurement using \$\$getrc3.bas (after Jos) on one port pin

```

Declare Sub Getrc3(w)
Config Lcdpin = Pin , Db4 = Portc.4 , Db5 = Portc.5 , Db6 = Portc.b ,
      Db7 = Portc.7 , Rs = Portd.7 , E = Portd.6
Config Lcd = 20 * 2
Dim W As Word

Do
  Call Getrc3(w)
 Cls
  Lcd W
  Waitms 300
Loop

Sub Getrc3(w)
Config Pind.3 = Output
Portd.3 = 0
Waitms 5
W = 0
Config Pind.3 = Input
Portd.3 = 0
Do
  If Pind.3 = 1 Then Exit Do
Else
  W = W + 1
  If W > 65534 Then Exit Do
End If
Loop
End Sub
    
```



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the BAS routine is 10 times faster than the original MCS routine GetRC0, because with $R_{max} = 10\text{ k}\Omega$ and $C = 100\text{ nF}$ a charging time of 5 ms (Waitms 5) is more than adequate.

Enter two port pins

The measurement will become a lot more accurate if the constant switching between input and output is avoided. This is made possible using the circuit in **Figure 2**, which 'occupies' two port pins. During the measurement, the capacitor is rapidly and fully charged via a low-value ($470\text{-}\Omega$) resistor on port line d.4. Next, pin d.4 is configured as an input, pin d.3 as an output and pulled Low.

The program consists of a clever subroutine which takes the place of $W = \text{GetRC9}(\text{pind},4)$

in the first listing. The subroutine (**Listing 3**) pulls both port pins High and charges the capacitor in 10 ms. Next, pin d.4 is returned to 'input' and from then on watches the capacitor voltage. The capacitor is allowed to

discharge all the time via port pin d.3. Software counter W is incremented on each loop iteration until the capacitor voltage has dropped below the Low limit. The loop is then left and the two port pins restored to the functions they had before the jump into the subroutine. Finally, there's the usual return instruction which takes the micro back to the main program.

The above measurement method enables errors of less than 1% to be achieved as long as the resistor value is within the range $1\text{ k}\Omega$ to $100\text{ k}\Omega$. Obviously, sacrificing an extra port pin is well worth the effort!

If you are low on processor I/O pins already then a usable measurement may also be carried out using the circuit in **Figure 3**, but only if the resistor to be measured is not too low in value. The trick used in the routine \$\$Getrc3.bas (Listing 4) is that the capacitor is charged in the

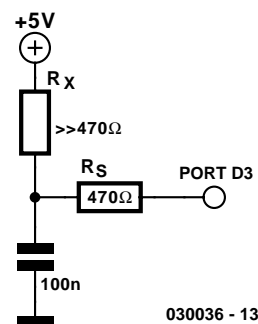


Figure 3. The port line saving alternative \$\$getrc3.bas.

'off' state and fully discharged via the port pin. The measurement sequence starts with port pin d.3 being configured as an input. The loop increments the counter until the voltage at the port pin reaches the High threshold value. Then follows the jump back to the main program.

(030036-1)

Minimalist Induction-Balance Metal Detector

A low-cost treasure hunter

Design by Thomas Scarborough

This design is arguably the simplest IB (induction balance) metal detector that could be built from off the shelf components. The IB method of metal detection has a good depth of penetration and discriminates well between ferrous and non-ferrous metals.



Metal detectors are available commercially but their price tag is often prohibitive to youngsters and those just starting out with the treasure hunting hobby. This article proposes a metal detector that is fun to build at a very low investment. Carefully built and aligned, it will clearly pick up a tiny 15 mm dia. brass coin at 70 mm in air, or a 25 mm coin at 120 mm. It will

not fail to pick these up vaguely at up to one quarter greater distance. The ability to locate coins buried in the soil are, of course, dependent on soil conditions, dry sand being the most favourable, and clay, the worst, 'medium' between the treasure and the search head.

The electronics

The electronics you'll need to build for this project is relatively simple and based on common or garden components only — some of which you may already have available in the junkbox or your regular component supply. The electronics (**Figure 1**) consists of a pulse transmitter and an associated receiver, with one-way traffic using two coils as a medium.

The transmitter is built around IC1. The CMOS (low-power) 555 generates a square wave output signal with a duty cycle of nearly 50% and a frequency of about 700 Hz. With the 555 in astable mode the output frequency is determined by components R4, R3 and C3. The output pulse is applied to TX (transmitter) coil L1 via series network R8-C4, with the electrolytic capacitor blocking the passage of DC through the coil, and the resistor protecting the output stage inside the 555. The pulse edges generated by the 555 will excite the coil and cause damped oscillation bursts at the self-resonance frequency of about 10 kHz.

The receiver section (IC2) is preceded by a simple yet effective pre-

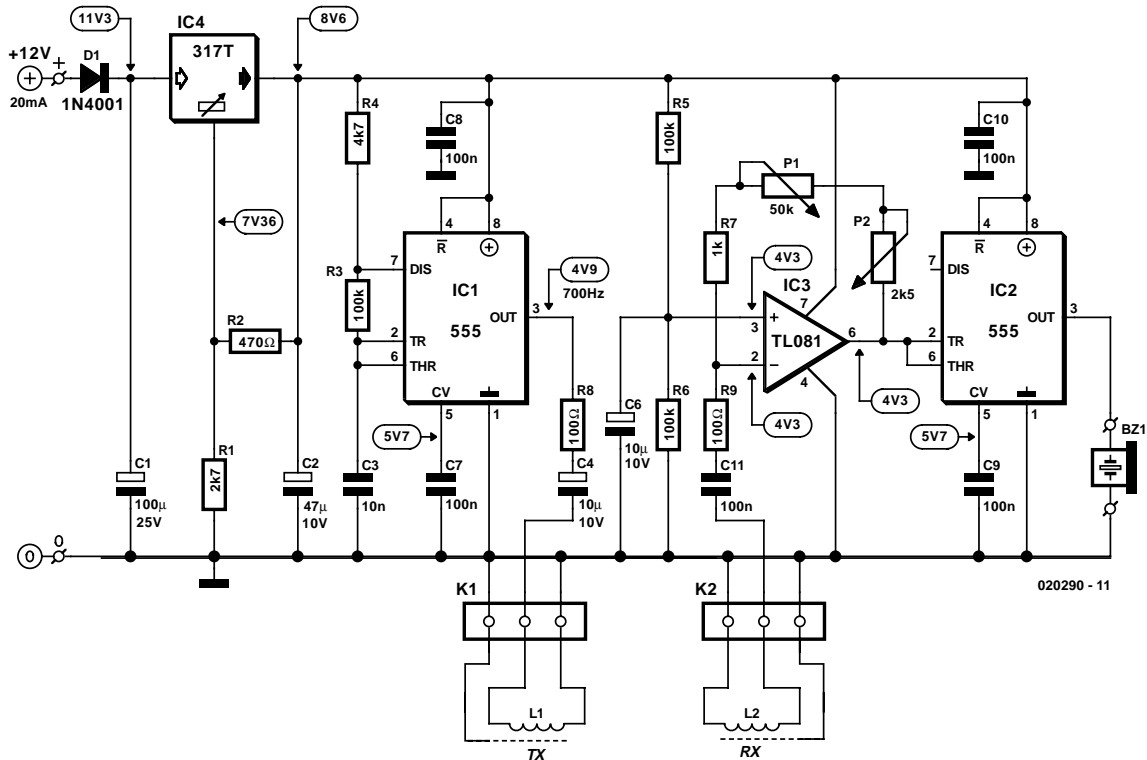


Figure 1. Circuit diagram of the metal detector.

amplifier stage based on opamp IC3, which amplifies the signal received from the RX (receiver) coil, L2 via C11-R9. The gain of the opamp is adjustable using P1 (coarse) and P2 (fine). The second CMOS 555 in the circuit, IC2, is configured to act as a true threshold detector, its output (pin 3) going High as soon as the signal level at the input (pin 2) drops below 1/3 of the supply voltage (or about 2.9 volts). Likewise the output drops low again as soon as the voltage at the THR input (pin 6) exceeds 2/3 of the supply voltage or about 7.4 volts). So, a 700-Hz signal will just start to sound from the piezo buzzer if the carefully adjusted threshold is exceeded. This adjustment is extremely critical and the 'crux' of the circuit.

Coupled coils

The TX and RX coils are critically coupled so that the presence of metal will disturb their magnetic coupling and with it the carefully adjusted 'quieting' of the threshold detector. Both coils have the same size and have a partial overlap. This enables the RX coil to pick up a positive as well as a negative (inverted)

portion of the magnetic field generated by the TX coil. Because at careful balance of the coils the positive and negative signals cancel out each other, in theory the RX coil will supply zero output signal. This we will call a 'null'. However, because of (fortunate) practical restrictions, a very small residual signal will always be generated. Once the delicate balance between the fields is disturbed by the presence of a metal object (which will absorb magnetic field energy) the RX coil will start to supply a higher output signal, causing the electrical threshold set up on IC2 to be exceeded, and the buzzer to start sounding. In practice, the detector adjustment is optimal with the **passive** piezo buzzer producing a soft crackling sound in the absence of metal. At this setting, the sound level increases considerably when a metal object is detected.

The adjustment of the 'null' position of the coils is critical and will be described further on.

The circuit is powered by a 12-V battery or a battery pack consisting of 8 penlight (AA) batteries or rechargeable cells. The use of an external battery pack carried on the shoulder will reduce the overall weight of the metal

detector considerably. It also enables a fresh pack to be substituted quickly.

The on-board voltage regulator, IC4, is configured with R1 and R2 to supply an output voltage of 8.6 V. Current consumption from the battery pack will be of the order of 20 mA depending mostly on the buzzer activity, of course.

Construction — the PCB

Populating the printed circuit board shown in **Figure 2** should not present problems because the board is spacious and only leaded components are used. Beginners should take strict guidance from the parts list and the component overlay printed on the board. As you can see from the photographs in **Figures 3 and 4**, the potentiometer shafts run through the board. This is done to allow knobs to be fitted once the shafts have been cut to a length which is determined by the enclosure. As we are dealing with a pretty sensitive piece of equipment, we recommend using a metal enclosure. This also allows the potentiometer cases to be grounded via the retaining nuts and washers.

Construction — search coil assembly

The constructing of the coils takes you into the much dreaded realms of 'mechanical con-

COMPONENTS LIST

Resistors:

- R1 = 2kΩ
- R2 = 470Ω
- R3,R5,R6 = 100kΩ
- R4 = 4kΩ
- R7 = 1kΩ
- R8,R9 = 100Ω
- P1 = 50kΩ potentiometer, linear
- P2 = 2kΩ potentiometer, linear

Capacitors:

- C1 = 100μF 25V radial
- C2 = 47μF 10V radial
- C3 = 10nF
- C4, C6 = 10μF 10V radial
- C7-C11 = 100nF

Semiconductors:

- D1 = 1N4001
- IC1,IC2 = 555C, TLC555, 7555 (CMOS)
- IC3 = TL071 CP, TL081 CP
- IC4 = (LM)317T (in TO220 case)

Miscellaneous:

- K1, K2 = 3-way PCB terminal block, lead pitch 5mm
- BZ1 = passive (ac) piezo buzzer
- PC1,PC2 = solder pins
- 12V battery or battery block (8 AA penlight cells)
- 36-30 SWG (0.2-0.3mm dia.) enamelled copper wire, 2 x 50m
- Enclosure: Bimbox 5004-14 or Hammond 1590B (109x58x25mm)
- 5m symmetrical screened microphone cable (two wires with common screening)
- PCB, order code **020290-1**

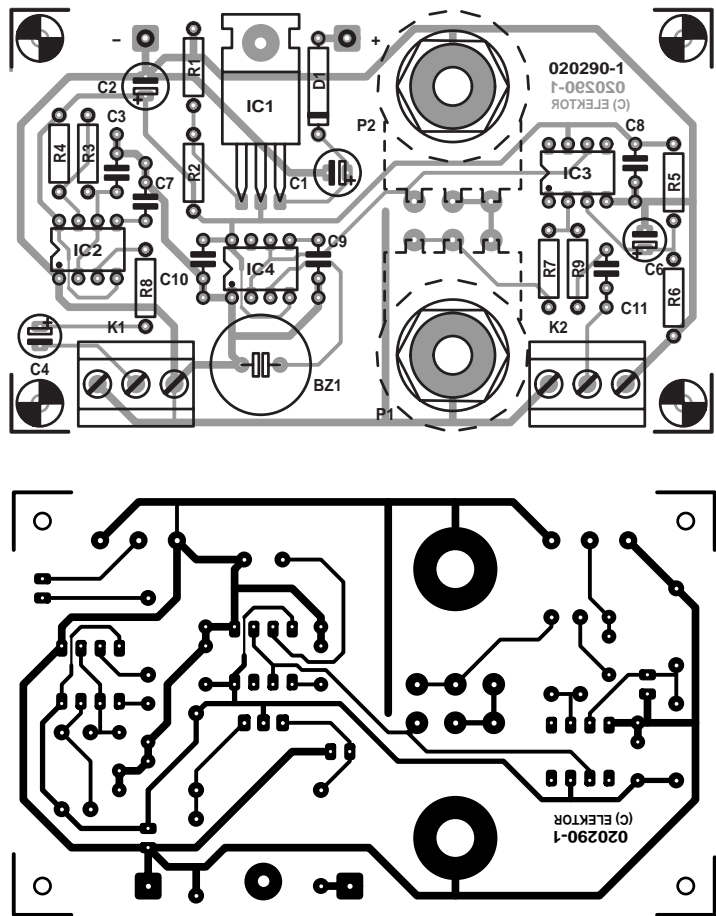


Figure 2. Circuit board design (board available ready-made).

struction' although some would argue for 'tinkering away on a Sunday afternoon'. Whatever, it will present a welcome change from handling your soldering iron and those components you know so well.

Below is a search head description (as suggested by the author) which we have no reason to question for veracity! Because a picture tells more than a thousand words, reference is made to **Figure 5**.

Both coils are identical. If you can get it, use 33 SWG (0.26 mm) enamelled copper wire, winding 100 turns in a clockwise direction on a circular 15 cm diameter former. The wire diameter is not critical — anything between 0.2 mm and 0.3 mm will do just fine. The coil is then temporarily held together with stubs of insulating tape passed under it and pressed together over the top. It may be jumble-wound. A second coil is wound in the same way. Each coil is then tightly bound by winding insulating tape around its entire circumference.

Next, add a Faraday shield to each coil. This is accomplished with some long, thin strips of aluminium- or tin-foil. First scrape the enamel off the base of each coil's end wire. Solder a 100 mm length of bare wire to the base, and twist this around the coil, over the insulating tape. This provides electrical contact for the Faraday shield. Beginning at the base of this lead, the foil is wound around the circumference of the coil, so that no insulating tape is still visible under the foil — but the foil should **not complete** a full 360 degrees. Leave a small gap — say 10 mm — so that the foil does not meet after having done most of the round (if you go all the way, you'll create short-circuited winding that will introduce a formidable amount of unwanted attenuation). Do this with both coils. Attach each of the coils to quality **balanced** (or 'symmetrical') screened microphone cable, with the Faraday shields being soldered to the screens of the cables. Do not use "stereo"

screened (microphone) cable — this may cause interference between the coils. The cable you need has two twisted signal wires inside a common screening braid. Each coil is now again tightly bound with insulating tape around its entire circumference. Last of all, wind strips of absorbent cloth around each coil (dishwashing cloth should suit), using a little all-purpose glue to keep them in place. Later, when resin is poured over the coils, this cloth meshes the coils into the resin. Gently bend the completed coils until each one is reasonably flat and circular, with each end wire facing away from you, and to the right of the beginning wire. Now bend them further, until they form lopsided ovals — like capital D's, as shown in the diagram. The backs of the D's are positioned so that they overlap each other slightly on the search head — this is the critical part of the operation, which will be performed after the printed circuit board has been completed.

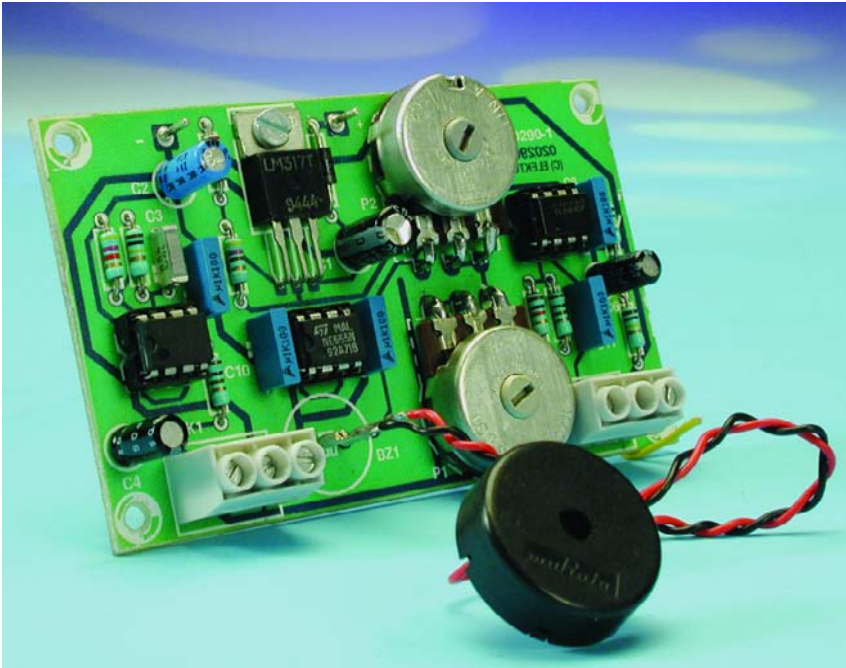


Figure 3. Top side of completed board.

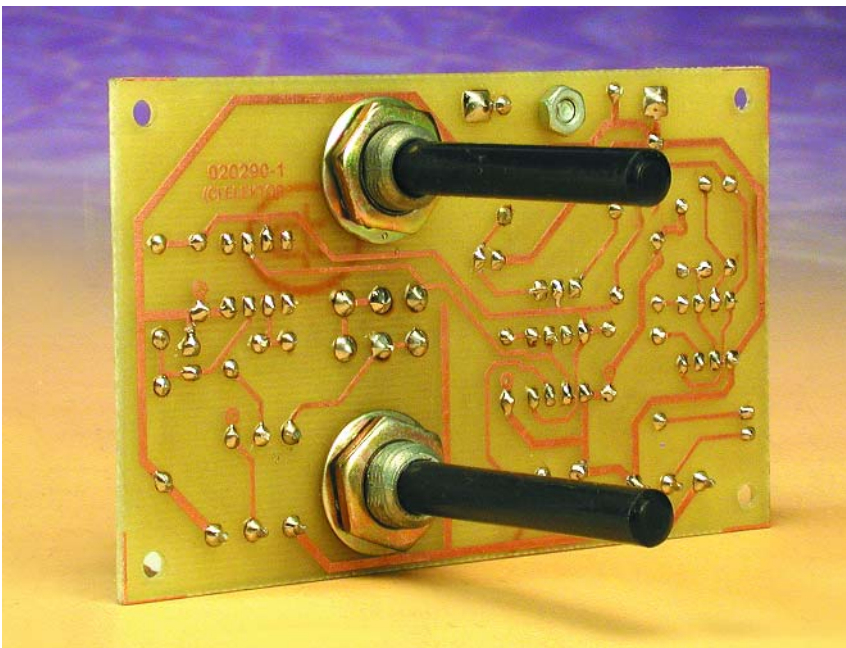


Figure 4. Underside of completed board.

Again, the coils need to be screened (with tin- or aluminium foil) to make sure they only respond to their own magnetic field, eliminating the risk of signal breakthrough (caused by even the smallest degree of capacitive coupling). Defective or inadequate screening will result in fast pulse edges from IC2 arriving directly at the input of IC3. These stray pulses will make it impossible

to find a 'null' in the search coil circuit (see below).

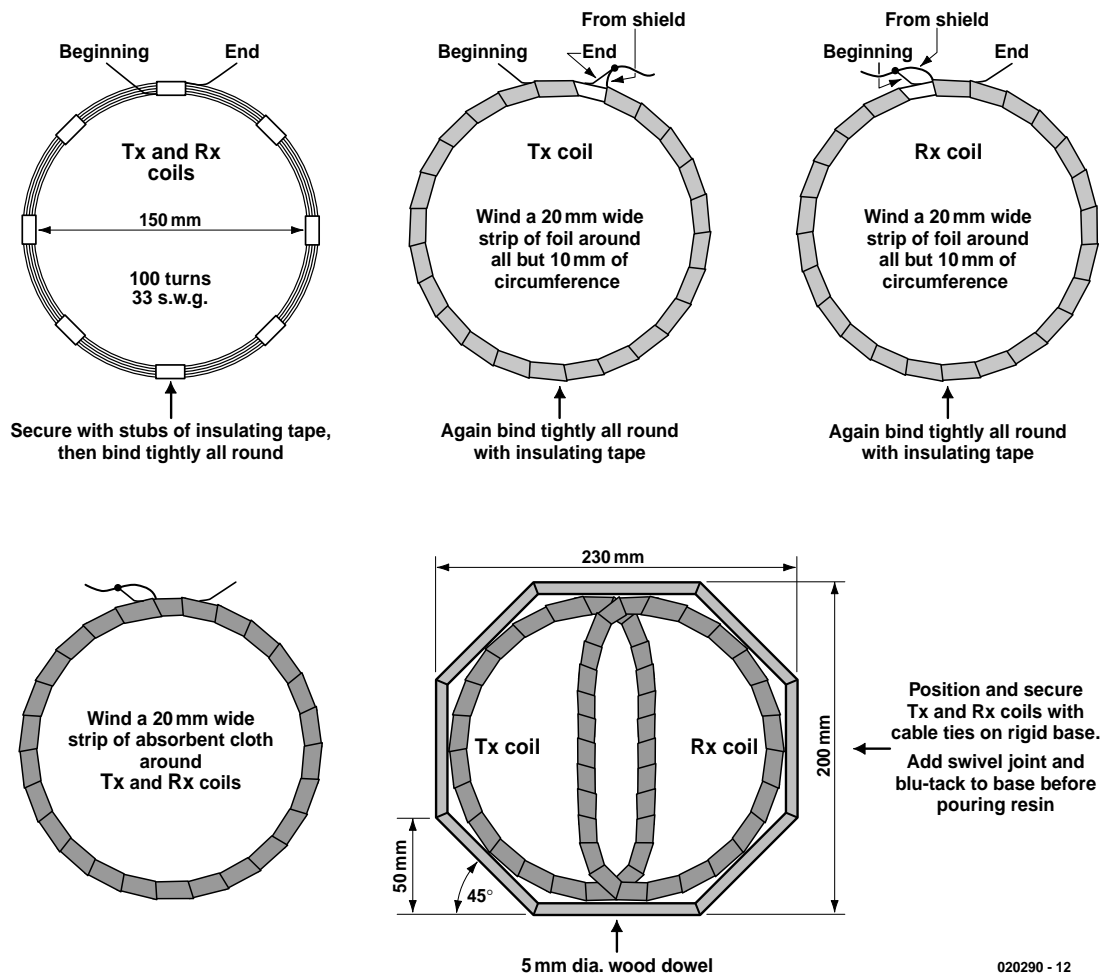
Adjustment

As you may have gathered, construction of the search head is not yet finished. At this point, however, some construction aspects start to interact with the adjustment of the circuit. Confused? Read on.

A completed and properly working printed circuit board is needed before we can 'pot' the coils. They are potted with plastic resin in a hard non-metallic base-plate. Any base-plate of suitable size will do, on condition that it is **rigid**. The author's prototype was potted on a suitably cut piece of masonite, with 'walls' made of 5 mm wood dowel glued around its edges. These walls should be 'resin-tight'. Do not use any metal in the search head. Also, cover a 4 cm segment of coil with Blu-Tack (Pres-Stik) so that it remains unpotted, and may be bent for final adjustment if required.

First place the coils on top of one another — ensuring that they are correctly orientated, with each end wire facing away from you, and to the right of the beginning wire. Turn the gain pots fully counter clockwise (minimum gain) to reduce the amplification to minimum. Attach a 12 V battery pack, and switch on. Now slowly move the coils apart until the buzzer is quiet. This is where the voltages in the Rx coil 'null'. Increase the gain somewhat and tweak the coil positions again. Repeat this adjustment several times, each time increasing the gain. The higher the gain can be set when still finding a null, the more sensitive the detector will be. **Always move the coils from fully meshed to fully apart, not the other way around.** If you adjust at the 'wrong side', the signal level on the coils will first drop when a metal object is detected, pass through the null, and then increase again to a level where the buzzer starts to sound. That's ok in principle, but the detector will be very insensitive.

Take your time to positively identify the optimum coil positions. If necessary use a wooden jig for easier mechanical adjustment and comparison of results. Once you've found the precise point where the coils need to be set, take a marker pen, and mark for holes around both coils. These holes are used to pass cable-ties through, to hold the coils tightly to the base-plate. Also use a cable-tie to hold the microphone cables to the base-plate. Use some Blu-tack (or Pres-stik) to tightly seal the holes underneath the plate before pouring the resin — plastic resin can be very 'runny', and sticks faster than many glues. Carefully bend the coils at the centre of the plate until you reach the exact balance at which there is neither silence nor screaming in the buzzer or headphones, but a crackle. A little drift should not matter at this point. You might also wish to attach a swivel-joint to the search head at this point, which will be attached later to the metal detector's shaft. One-inch (or larger) PVC kitchen drainage conduit may be used for the shaft. Now you are ready to mix and pour the resin.



020290 - 12

Figure 5. Search head construction guide.

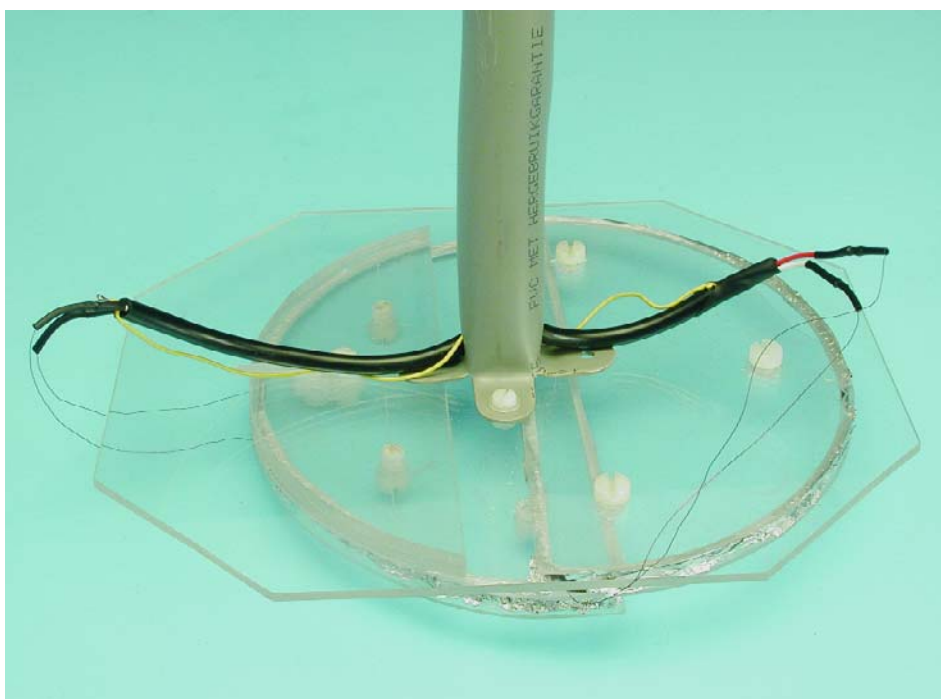


Figure 6. Alternative search head assembly made from Perspex.

Use the correct amount of catalyst, so that there is not too much heat and shrinkage in the resin. Pour the resin over the cloth which surrounds the coils, so as to soak it, and keep on pouring at least until the entire bottom of the plate is covered with resin. The circuit may no longer function correctly at this point until the resin has hardened, so make no more adjustments at this stage, but switch the circuit off.

When the resin has set hard, keep the search head away from all metal, and away from computer equipment, which can cause serious interference, and switch on. Adjust potentiometer P2 (fine-tune) to their mid-points. Then adjust P1 (tune) until the metal detector is just at the point where a crackle is heard, between silence and a scream. Use the tune and fine-tune knobs for any further tuning. Move a coin over the search head, and the piezo sounder should 'scream'.



Figure 7. Milling the grooves in the Perspex semicircles.

Alternative construction method

Figures 6 and 7 show an alternative construction method for the search

head, based on the use of pieces of Perspex. The TX and RX coils are tucked away in a groove milled in the side of each semicircle of Perspex. Another piece of Perspex, this

time square shaped, is used to hold the semicircles together, secure them to the PVC pipe, and still allow the semicircles to be adjusted. The latter is achieved with the aid of nylon screws. This assembly, when finished and mechanically adjusted, should still be cast in resin or securely glued to warrant the necessary ruggedness.

In actual use

You'll soon find that the adjustment of the metal detector will be affected by the mineralisation of the ground you are searching, as well as temperature and voltage variations — so readjustments to P1 and P2 are inevitable. One trade-off for the extreme simplicity of this design is some drift. While it is not excessive, re-tuning will be necessary on a fairly regular basis.

At the centre of the search head, the circuit's rejection of iron is very high, so that once you have a 'feel' for the detector, iron can be virtually excluded. This is a boon for anyone who is searching in the first instance for coins or noble metals.

(020290-1)

Low-Cost Inductance Meter

July/August 2003, p. 58,
034030-1

The formula to determine the coil inductance should read: $L = 1.2 \, dU/dt$, not $L = 120 \, dU/dt$.

The error in the calculation starts at: $dI_L/dt = 0.1 \, dU/dt$; this should be $dI_L/dt = dU/dt / 0.1$.

EEDTs Pro Budget Booster

July/August 2003, p. 43-44,
014098-1

The type numbers BDY66C (T3) and BDV64C (T4) printed in the parts list and the circuit diagram are wrong and should be replaced by:

T3 = BDV64

T4 = BDV65

The alternative type num-

bers BDW84 for T3 and BDW83 for T4 are correct.

Week/Month Timer

April 2003, p. 66-69,
020190-1

One of our readers kindly informed us that a number of ICs type CD4068 or HCF4068, including those manufactured by Philips Semiconductors, have an

additional output on pin 1. The ST Microelectronics HCF4068 used in our prototype did have such an output.

Part of the rule mentioned under Step 2 in the section 'How to set the period' is incorrect and should be amended to read: '... the first number that's smaller than or equal to the above value.' This rule is, in fact, applied in the example that follows.

Valves at Low Plate Voltages (2)

Part 2: more power!

by B. Kainka

Genuine power valves such as the EL84, EL95, ECL80 and ECL86, and in particular the PL504, do of course offer more power output at low anode voltages than the ECC81 and ECC82 types we discussed in the first part of this series. With the PL504 it is even possible to drive an ordinary low-impedance loudspeaker, using an anode voltage of only 27 V. In this second article we will also be looking at some miniature Russian 'battery valves', which not only operate at low anode voltages, but also demand considerably less heater power.

If, after our initial experiments with ECC81s and ECC82s (or their US equivalents 12AT7 and 12AU7), you have come to the conclusion that our valve headphone amplifier could do with a little more power, then it is time to take a look at some valves which are designed for higher power. Suitable candidates would be output-stage valves such as the EL84 (6BQ5), EL95 (6DL5), ECL80 (6AB8), ECL86 and similar types.

The EL95

Here we tested a used EL95. The EL95 is a pentode, which means that it has two more grids than a triode. Grid 2 should be connected to the supply voltage, while grid 3 should be connected to ground. Pentodes have a higher gain and lower distortion than triodes. The EL95 is used here with an anode voltage of between 12 V and 24 V. **Figure 1** shows the circuit arrangement for one channel and the pinout of the valve.

The EL95 is relatively economical, drawing a heater current of only 200 mA. Despite

this, we obtain an anode current of 1.3 mA at 12 V and 3.5 mA at 24 V. A respectable power output can be achieved using a suitable transformer. A usable stereo headphone amplifier for low-impedance (32 Ω)

headphones can be constructed to run from a 12 V supply, for example from an ordinary mains adaptor. The heaters of the two valves should be wired in series to the same 12 V supply used for the anodes.



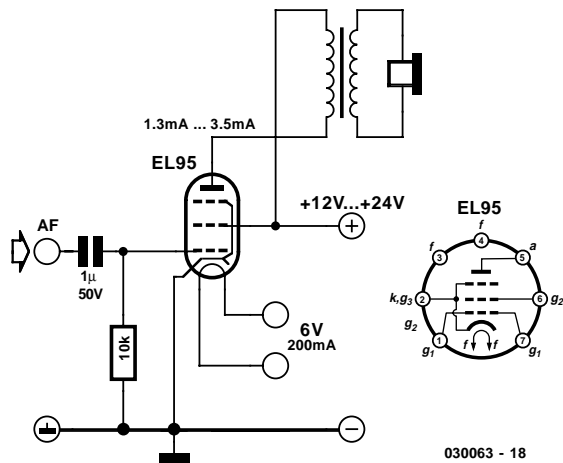


Figure 1. Amplifier using an EL95.

If good sound quality is important then the right transformer must be used. Very small mains transformers have a high winding resistance and saturate too easily. If larger transformers are used, the saturation problem disappears, but the inductance will be lower and so low frequency response will be impaired. A mains transformer can only replace a genuine audio transformer in limited circumstances: in general a custom transformer must be wound for each design of valve output stage, which is of course rather inconvenient.

And so we return to the simplest design and connect the headphones directly in the anode circuit. Experiment shows that this works well with 600 Ω headphones. The question is whether the headphones can tolerate a direct current of 3.5 mA, for example. This corresponds to a

static dissipation of just under 7.5 mW in the quiescent state. The thermal effect of this does not seem excessive, since headphones are typically rated up to 100 mW. A different question is whether the mechanical displacement of the diaphragm will lead to distortion. The direct current is effectively a mechanical pre-tensioning, or 'bias', of the diaphragm, which could, in principle, affect the sound. In practice, however, there is no perceptible disadvantage, and so it seems a sensible idea to construct a headphone amplifier without transformers.

If it is desired to avoid both the use of transformers and a direct current in the headphones, then capacitor coupling can be used (Figure 2). In this case, high-impedance (600 Ω) headphones must be used. The volume is lower than with an ideal out-

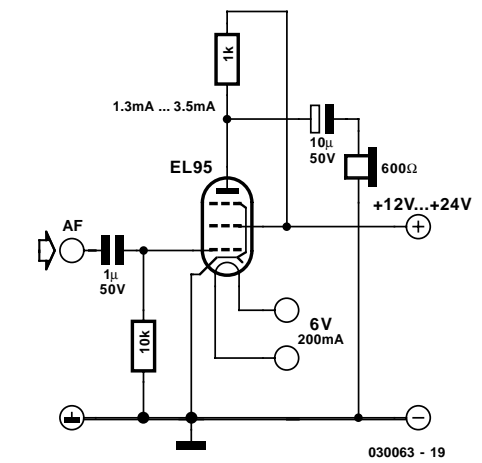


Figure 2. Transformerless headphone amplifier using an EL95.

put transformer, but is entirely adequate for most applications.

Class A amplifier using a PL504

Can we do a little more? What about using a PL504 in a small class A loudspeaker amplifier? The PL504 is a physically large valve with a Magnoval base, which was used in the line output stages of television sets. Its heater requires 300 mA at 27 V. It therefore makes sense to use an anode voltage of 27 V also. In its intended application in the line output stage the valve had to operate at anode currents in excess of 500 mA, and of course this means that we can also expect good performance at lower voltages.

An experiment using 27 V produces an anode current of 33 mA. That is already more than (according to its data sheet) the EL95 can deliver at 250 V — we will certainly be able to build a very useful loudspeaker amplifier using a PL504 at 27 V. A reasonably large transformer will be needed for good results. The impedance connected to the valve should be around 800 Ω. The circuit was tested using a large mains transformer with a 230 V primary and a 24 V secondary, having a turns ratio of around 10:1 and hence an impedance ratio of around 100:1. An 8 Ω loudspeaker is then just right, since the valve will see 800 Ω. An oscilloscope can be used to check that the valve is suitable. If, when the amplifier is driven into limiting both positive and negative half-cycles are clipped to an approximately equal degree, then the circuit is optimally arranged.

In use, this amplifier (Figure 3) delivers plenty of volume and a pleasant sound. The anode power dissipation is around 1 W, as would be expected. The 27 V supply voltage is a little inconvenient: an experiment was tried using 24 V. The lower heater power does

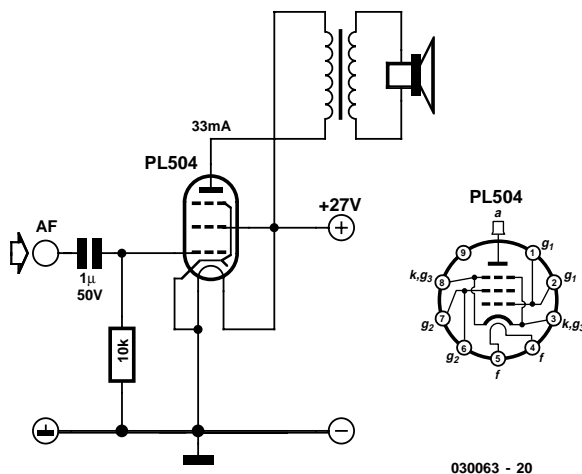


Figure 3. Loudspeaker amplifier using a PL504.

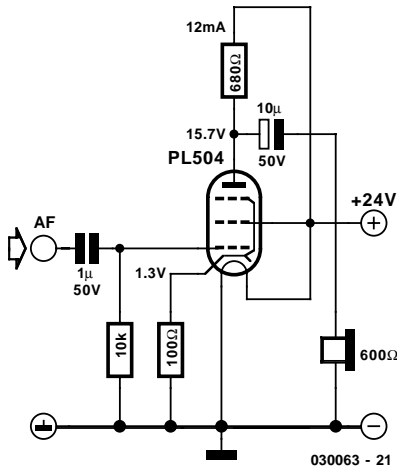


Figure 4. Headphone output stage running on 24 V.

not appear to have any ill effect; the anode current falls to 25 mA and there is practically no effect on the sound.

PL504 headphone amplifier

When building a headphone amplifier, the PL504 has plenty of spare capacity to allow RC coupling to be used at the output. A cathode resistor of 100 Ω provides a grid bias voltage of -1.3 V. At the same time, distortion is reduced due to negative feedback. A 680 Ω resistor in the anode connection allows the amplified signal to be coupled out of the circuit. In this circuit (Figure 4) only about half of the amplified signal current flows through the headphones, the other half flowing through the anode resistor. With an anode current of 12 mA we can put up with this loss.

This headphone amplifier can be constructed without difficulty, since there is no need to find a suitable output transformer. The circuit has a pleasant sound and plenty of volume. If you can find two PL504s, consider yourself lucky; alternatively, there is the EL504 which has a heater voltage of 6.3 V at a current of 1.3 A. This is practically the same valve, but with a different heater. And of course there are many other power pentodes around which similar circuits can be designed.

Low-voltage valves from Russia

The 1SH24B, 1SH29B and 1P24B (Figure 5) are tiny battery valves from the former Soviet Union. These valves are still available in large quantities and can be obtained very cheaply. They have wire connections that can be soldered directly, and so special sockets are not required. A characteristic of these valves is the directly heated cathode: in other words,

the heater is also the cathode. This has an impact on the circuit design, since it makes it difficult to wire heaters in series.

Most valves have a concentric construction, but these are rather different. In the centre is a thin heater element which forms the directly-heated cathode; in the case of the 1SH19B and 1P24B there are two cathodes. All the other electrodes take the form of plates or wires arranged parallel to the cathode. This makes for a very robust and efficient valve.

The 1SH24B (high frequency pentode) has a heater current of just 13 mA at 1.2 V — a marvel of efficiency. The 1SH29B (universal pentode with $P_a = 1.2$ W) has a heater current of 64 mA at 1.2 V or 32 mA at 2.4 V.

Initial experiments with the 1SH29B show that it has very useful properties: at $U_a = U_{g2} = 40$ V the

anode current is 3 mA, grid 1 being held at the same voltage as the negative heater terminal. The transconductance is about 1 mA/V. The valve also works at lower voltages, but the anode current and transconductance fall off sharply. Even at an anode voltage of 12 V the characteristics are better than those of the ECC81.

Still better is the 1P24B (power pentode with $P_a = 4$ W). The 'P' might stand for 'power', and this tiny valve certainly has that. It needs a relatively high heater current of around 240 mA at 1.2 V, but in return we get very useful results even at an anode voltage of zero. With a grid voltage of zero we obtain an anode current of 2 mA and a transconductance of 1.5 mA/V. The valve is therefore particularly suitable for use in a small headphone amplifier. Figure 6 shows a tested circuit for use with high-impedance headphones.

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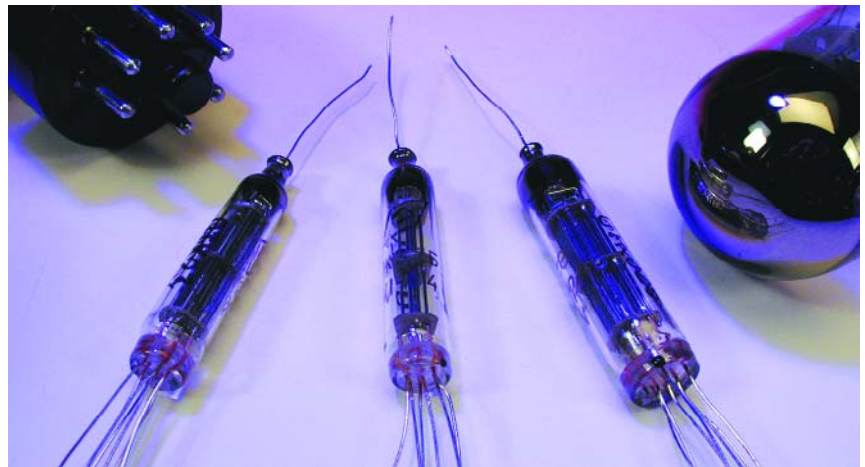


Figure 5. The 1SH24B, 1SH29B and 1P24B.

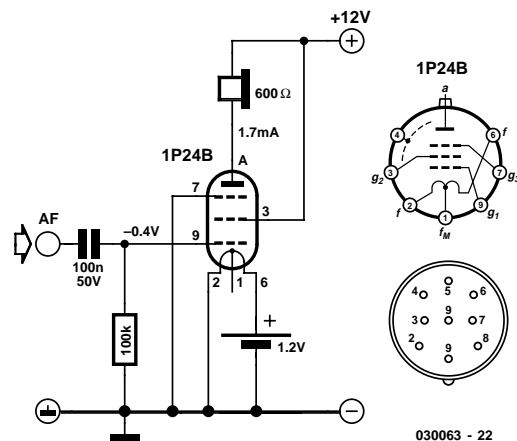
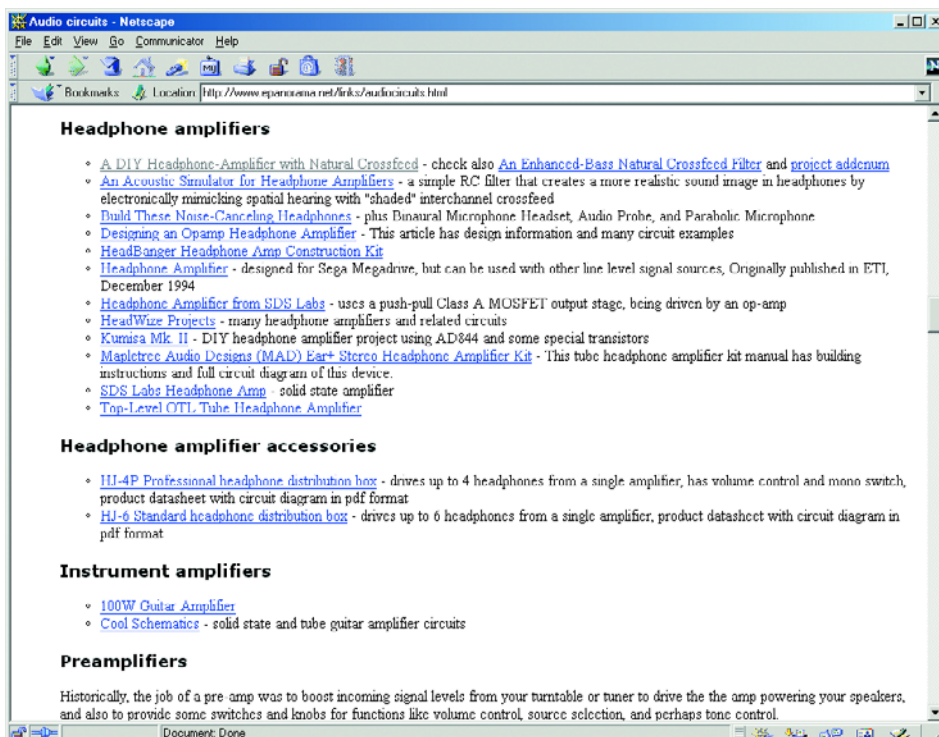


Figure 6. Audio amplifier using a 1P24B.

Headphone Amplifiers

DIY with valves and semiconductors

By Rolf Gerstendorf



Most commercial pre-amplifiers and power amplifiers have a headphones output. However, because of its simplicity the sonic quality of such an output is often judged as 'mediocre'. Optimum sound reproduction in a pair of expensive headphones requires a dedicated amplifier and these, unfortunately, are rarely seen in the highstreet hi-fi outlets. The Internet however offers many alternatives provided you are prepared to go for home construction.

Many professional electronic engineers once started their 'career' with the construction of a headphone amplifier. This is not surprising, the circuits being relatively simple and requiring little theoretical knowledge of electronics. If you start trawling the Internet for circuit suggestions

and feedback on published designs, it is harder to separate wheat from chaff than to find anything at all.

At the very lowest level the circuits usually have either a discrete amplifier or an opamp acting as an impedance converter with a volume control in front of it. The range continues, however, with far more extensive designs right up to high-end projects for electrostatic headphones. Surprisingly, friends of the valve have contributed a lot to the DIY headphone amplifier scene on the Internet.

An excellent place to start your search is Chu Moys **Headwise** website [1] which is home to many specialized forums, FAQ lists, links and so on. In the DIY section called **Projects** you will find nearly 100 (!) project suggestions for headphone amplifiers of various complexity and with documentation ranging from meagre to sumptuous.

Cable manufacturer RJM employs the DIY section of its website **RJM Audio Cable** [2] to present DIY headphone amplifiers and modifications to Headwise designs.

The **Black Widow Audio** website [3] supplies two construction projects. The circuit versions developed for the 'crossfeed' principle are well worth looking at.

Further interesting sites we came across are **Audio Pages** [4] and **Audiologica** [5]. The latter contains information on the META42 and MINT designs.

The **Wheatfield Audio** website at [6] has a surprise in store in that a commercial product is published for DIY construction, complete with schematics and PCB layouts.

Searching the **Epanorama** website at [7] for the argument 'headphone' produces a lot of links, not only to projects for home construction, but also two articles covering various subjects like headphone amplifier simulation, calculation tools and reviews of commercial kits. A similar collection of links is found on the **Innehall** site at [8].

On his website at [9] **Raymond van Weeghel** presents, among others, a headphone amplifier in hybrid technology. A valve is used in the prestage and a MOSFET in the output stage.

(035060-1)

A Pocket Headphone Amplifier

by Chu Moy

"Thank you for your amplifier design. I built it and can't believe how wonderful it makes my AKG K340 headphones sound as well as my Sennheiser 600."
- A DIYer.

While doing research for the article [Designing an Opamp Headphone Amplifier](#), I built a portable headphone amplifier for testing purposes. Each channel uses a single Burr-Brown OPA134 opamp in a non-inverting configuration. It has adequate current capability to drive most headphones without an output stage. I have used it with Sennheiser 465s (94dB SPL) and achieved ear-splitting volume. The amplifier is ideal as a booster for power-conserving stereo sources such as portable CD players and for interfacing with passive EQ networks such as tone controls or a [headphone acoustic simulator](#).

The Amplifier Design

C1 - 0.1uF, 100V capacitor
R4 - 10k resist, 1/4W resistor

Web addresses:

- [1] http://headwise2.powerpill.org/projects/showproj.php?file=meier_prj.htm
- [2] www.geocities.com/rjm003.geo/rjmaudio/diy_hclo.html
- [3] www.blackwidowaudio.com/index.htm
- [4] <http://sound.westhost.com/projects-1.htm>
- [5] <http://tangentsoft.net/audio/>
- [6] www.pmllett.addr.com/ha-4_headphone_amp.htm
- [7] www.epanorama.net/links/audiocircuits.html
- [8] http://audionova.nu/Innehall/audionova_DIY_fr.htm
- [9] www.raymondaudio.nl/