500 ppm* LCR Meter (1)

The luxury of precision within everyone's reach

By
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The remarkable precision of this device and its amazing ease of use are the result of careful design. It works so well behind its uncluttered front panel that one could almost forget the subtleties of the measurement techniques employed. A dream opportunity for our readers who are passionate about measurement to enjoy themselves. If, like us, you wonder at the marvels modern techniques bring within our reach, come along and feel the tiny fraction of a volt.

Apologies

It is impossible to tell the full story on this test instrument in one go, so we'll just have to accept right away the idea of breaking it up into manageable chunks. You will only have seen the whole thing once the series of two or three instalments have been published — please excuse this inconvenience.

* *see detailed specifications*

Technical specifications			
Display	Dominant parameter: value Secondary parameter: value Equivalent circuit: series or parallel (manual or automatic selection) *Q*-*D* (can be reversed with respect to the automatic choice) $\|Z\|$ Φ either *R*s + *X*s, *V*x + *I*x, or ADCU + ADCI		
Sort function	On the main parameter of a control component (*R*, *L*, or *C*) after adjustment of the central value. Tolerances: 1 % 2 % 5 % 10 % 20 %		
Measurement scope	**Parameter**		**Value**
	L		0.1 nH – 100 H
	C		0.1 pF – 100 mF
	R, $\|Z\|$		0.1 mΩ – 1,000 MΩ (1GΩ)
	Q or *D*		0 – 10,000
	Φ		−90.00° to +90.00°
	R_s, X_s		0.1 mΩ – 1,000 MΩ (1 GΩ)
	U_x and I_x	U_x	0 – 500 mV
		I_x	0 – 5 mA
	ADC *U* and ADC *I*		0 – 5 V
Test frequencies	50 Hz supply		100 Hz, 1 kHz, 10 kHz
	60 Hz supply		120 Hz, 1 kHz, 10 kHz
Power consumption	Without display		5 V – 100 mA
	With backlit display		5 V – 180 mA
PC software	For Windows, Linux, MacOSX		

Test conditions	
Off-load test voltage	0.4 V_{rms} ± 5 %
Ranges	8, automatic
Measuring speed	Around 2 measurements per second. It is possible to take the average of several measurements (1–9), at the expense of speed. A green LED indicates the end of each sequence.

Accuracy of the main parameter (*R, L, C*)	
Conditions*	Warmed up for 10 minutes, 25°C ± 2°C Use of 0.01 % resistors (100 Ω, 1 kΩ, 10 kΩ, 100 kΩ) in the current/voltage converter.
Ranges 3, 4, 5, and 6	< ±500 ppm (0.05 %) (down to 200 ppm **) ±1 on the final digit
Ranges 2 and 7	< ± 1000 ppm (0.1 %) (down to 800 ppm) **) ±1 on the final digit
Ranges 1 and 8	< ± 3000 ppm (0.3 %) ±1 on the final digit
** The accuracy (±1 unit in the last figure on the right) is greatest when the post-amplification ranges are the same for *U* and *I*.	

Miscellaneous	
Measurement connections	4 Kelvin wires on BNC connectors
Compensations	"OPEN" / "SHORT"
"SHORT" limit	$\|Z_s\| < 10\ \Omega$
"OPEN" limit	$\|Z_s\| > 100\ k\Omega$
Supply voltage	5 V_{DC} ± 5 %, via the USB connector

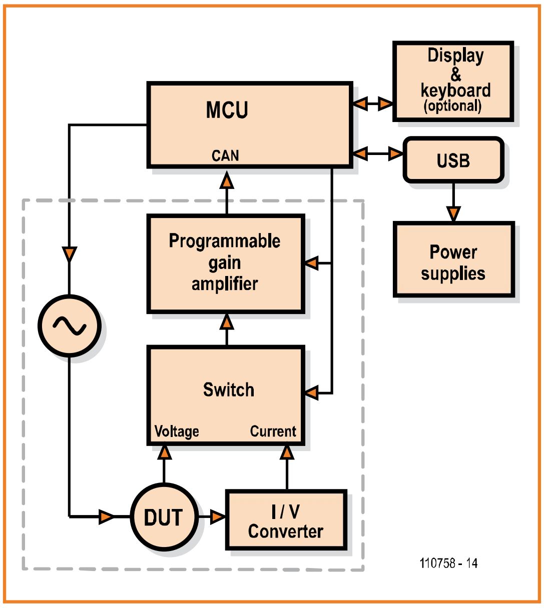


Figure 1.
Analogue and digital techniques are closely intertwined within the sub-assemblies of the 500 ppm LCR Meter.

An LCR meter isn't usually seen as indispensable in an amateur electronic enthusiast's workshop. However, with the proliferation of SMD components with no markings, like chip capacitors, or the increasingly common use of chokes in switch-mode PSUs, the use of an LCR meter is becoming more and more frequent. Let's not forget too that this device doesn't stop at just giving a passive component's 'value' in the usual sense — inductance *L*, capacitance *C*, or resistance *R* — but also its secondary component, defining its 'quality' (see specifications box), which can be defined in several ways:

- Φ – phase angle between voltage and current: $\tan \Phi = |X_s| / R_s$
- *Q* – quality factor = tan Φ, used to characterize an inductor.
- *D* – dissipation factor = 1/*Q*, used to characterize a capacitor.

This duplicity in our components, innocent enough as to be most usually ignored in low-frequency circuits, does need to be taken into account in high-frequency work, and more generally in precision circuits.

We have to go back more than 15 years in the Elektor archives to find a precision LCR meter [1]. My project, the fruit of which you're going to be discovering here, has itself been in gestation for four years; the first version, battery-operated, with a modest 2×16 LC display, has become a bench model, mains powered, with a 128×64 graphic display. My exchanges with Elektor Labs have led to the version published here, which benefits from all the accumulated experience. Did we have to choose between two concepts, either a computer used as a display and control peripheral (USB link) associated with an LCR meter that was reduced to being just a measuring interface, or a true self-contained LCR meter? No. To keep everyone happy, I'm proposing a no-compromise variable configuration, designed to combine accuracy and convenience at the highest level:

- A measuring head that dialogues with a PC (for displaying the results and sending commands), but is also present on an extension connector;
- An optional extension comprising a display and a mini keyboard, which when connected to the previous element converts it into a **standalone** unit;

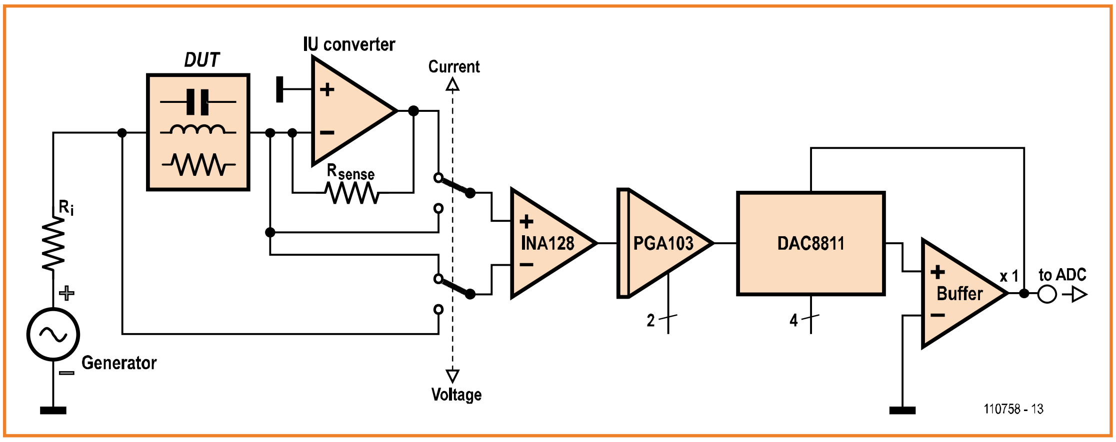


Figure 2.
An enlargement of the section within the dotted line in Figure 1: this is the measuring head, which most of this article is about.

A little bit of theory

The complex impedance to be measured is equal to the ratio between the vectorial dimensions $\overrightarrow{U_X}$ and $\overrightarrow{I_X}$, representing the voltage across the device under test (DUT) and the current flowing through it:

$$\overline{Z_X} = \frac{\overline{U_X}}{\overline{I_X}}$$

Each vector can be broken down into *phase* and *quadrature* components with respect to some fixed reference:

$$Z_X = \frac{V_p + jV_q}{I_p + jI_q}$$

Hence again, using the *series* representation of an impedance $Z_x = R_S + j\,X_S$

$$R_S = \frac{V_p\,I_p + V_q\,I_q}{I_{p^2} + I_{q^2}} \qquad X_S = \frac{V_q\,I_p - V_p\,I_q}{I_{p^2} + I_{q^2}}$$

Some LCR meters go down the analogue route (phase detectors) to obtain the phase and quadrature components of the voltage and current to be measured, the final measurement being performed by an ADC, often of the dual-ramp type for good accuracy, as the DC voltages to be measured are in fact ‘contaminated’ by a not-inconsiderable residual voltage, if one is looking for a fast response time. The “all digital” method does not suffer from this drawback, and the mathematical operation of discrete Fourier transformation (DFT) makes it possible to obtain the *phase* and *quadrature* values for the voltage (U_p U_q) and current (I_p I_q) from N samples d_i of one period of the signal to be measured

$$U_p = \frac{1}{N}\sum_{i=0}^{N-1} d_i \times \cos\left(\frac{2\pi i}{N}\right) \qquad U_q = \frac{1}{N}\sum_{i=0}^{N-1} d_i \times \sin\left(\frac{2\pi i}{N}\right)$$

This requires just a fast, accurate ADC, and a little bit of calculating power.

- Software, within the measuring head, capable of handling both these modes.

Before we take a detailed look at the complete circuit of the unit, I propose first taking a quick look at the block diagram, which gives a good idea of the whole project (**Figure 1**). It would be tempting to immediately examine each element in detail as we usually do. Sadly, we wouldn’t understand a great deal of it without first taking a look at the problems posed by these measurements. All in all, they are fairly complex, and the precision requirement we’ve set for ourselves places the bar pretty high. Rest assured, though, that all I’m using for this is grey matter — there are no other hard-to-find components.

Principle and operation

In the “A little bit of theory” box, I provide a few details about the measurement principle. Let’s take a look here at its practical implementation in the measuring head (**Figure 2**), which comprises those elements enclosed by the dotted line in Figure 1. The analogue section of the LCR meter uses the conventional technique of a self-balancing bridge to determine the unknown impedance of the device under test (DUT) by measuring the voltage across its terminals together with the current flowing through it when driven by a variable-frequency sinewave signal.

We can see from Figure 2 that the current flowing through the DUT also flows through the current sensing resistor R_{sense} (this simplified diagram doesn’t show that the value of this sensing resistor changes with the impedance range, which we’ll see in the full circuit diagram). The potential at the inverting input of the current/voltage converter (*IU converter*) is maintained at 0 V (virtual ground) so as to maintain the balance between the current through R_{sense} and that through the DUT. As the frequency is never more than 10 kHz, a fast opamp is suitable for this converter, which must introduce only a minimum of phase error into the signal path.

The complex impedance of our unknown component will thus be obtained from the voltage measured across the DUT and that across R_{sense} (an image of the current through the DUT), applied to a differential amplifier (INA128) via the I/U switch. Before being digitized by the microcontroller, the signal undergoes further amplification (PGA103), multiplication (by means of the DAC8811 fast converter), and filtering. The remainder of its path is handled by the software, which will first determine the basic series parameters of the unknown impedance R_s and X_s (where X_s is an inductive or capacitive component, depending on the nature of the DUT) and then lastly the

Table 1. Combined gain: variable from 1 to 866 in 48 steps															
PGA gain	1					10					100				
CNA gain	1	1.155	..	7.50	8.66	1	1.155	..	7.50	8.66	1	1.155	..	7.50	8.66
Total gain	***1***	***1.155***	***..***	***7.50***	***8.66***	***10***	***11.55***	***..***	***75.0***	***86.6***	***100***	***115.5***	***..***	***750***	***866***

Table 2. Measuring Ranges

Range	IU converter resistor	U gain	I gain	Measuring Range (resistance)	Measuring Range (L or C - impedance)
1	100 Ω	100	1	< 0.1 Ω	< 1 Ω
2	100 Ω	10	1	0.10 Ω to 11 Ω	1 Ω to 10 Ω
3	100 Ω	1	1	11 Ω to 900 Ω	10 Ω to 995 Ω
4	1 kΩ	1	1	900 Ω to 9.9 kΩ	996 Ω to 10 kΩ
5	10 kΩ	1	1	9.9 kΩ to 99.9 kΩ	10 kΩ to 100 kΩ
6	100 kΩ	1	1	99.9 kΩ to 1 MΩ	100 kΩ to 1 MΩ
7	100 kΩ	1	10	1 MΩ to 10 MΩ	1 MΩ to 10 MΩ
8	100 kΩ	1	100	> 10 MΩ	> 10 MΩ

other parameters derived by calculation: *Z*, *L*, *C*, *R*, *Φ*, *Q*, *D*.

The conformity of the successive stages is determining. In order to avoid problems of drift, the same chain is used for both the voltage and current measurements after the *IU converter*. The high precision of the programmable gain amplifiers used and the compensation for the stray differential phase error, which is a function of the gain of the chain, ensure a basic accuracy practically equal to that of the precision resistors used in the current/voltage converter. You could be forgiven if you now rushed ahead to look at the circuit, without reading what follows. However, do be aware that in order to understand everything, you will have to come back to the next two essential paragraphs, which may at first reading seem a bit hard going...

Better watch that gain

To obtain extended measuring ranges (see paragraph "Measuring ranges" below), the amplitude of the signals to be measured must be adjusted *before* they are digitized. This is achieved by:

- Selecting the appropriate value of R_{sense}, according to the impedance of the DUT. The values adopted are: 100 Ω, 1 kΩ, 10 kΩ, and 100 kΩ (we'll see these later on the circuit diagram).
- Modifying the gain of the measuring chain so as to drive the analogue/digital converter (ADC) with a voltage that's as high as possible, without overloading it. Using the same amplifying chain for the current and voltage measurements makes it possible to avoid a good part of the drifts and uncertainties about the global gain value. In point of fact, as the impedance value can be written:

$$Z_x = \frac{V_p + jV_q}{I_p + jI_q} \times \frac{Gi\ R_{sense}}{Gv}$$

Where

G_i and G_v are the current and voltage gains of the amplifier channel;

R_{sense} is the *IU converter* resistor;

V and I are the voltages measured by the ADC;

the ratio G_i/G_v will only keep the variable parts of the gains.

The linearity error of a successive approximation ADC (SAR) is at best ±1–2 LSBs. Now since the zero-crossing of the sinewave signal to be measured is the region where the digitizing error is greatest, the higher the amplitude of the signal, the more accurate will be the measurement. Most LCR meters, like for example the one described in *Elektor* in 1997 [1], use a programmable gain amplifier (PGA in Figure 1) using a step of ×10 between values (gain of 1, 10, or 100). It is quite simple to use a high-precision integrated PGA, such as the PGA103 from Texas Instruments.

Compensating the phase shifts in the *IU converter* and PGA103

The software measures reference capacitors

The phase shifts introduced by the *IU converter* are compensated for **by the software**. The phase shift to be compensated is determined by measuring components whose behaviour is known, such as SMD capacitors with NP0 dielectric, considered as 'perfect' (at these low frequencies). The advantage of this method: the stray capacitance of the connections is in parallel to the capacitance proper and hence is no longer of any importance.
We will be aiming for a −90° phase shift.
To do this under the best conditions, the amplification gains must be identical for *U* and *I*; this implies that @ 10 kHz the impedance of this DUT must be practically equal to R_{sense}: 159 nF for 100 Ω, 15.9 nF for 1 kΩ, etc.
To make things simpler, these capacitors are wired onto the PCB, but they are only brought into service manually via a jumper during the compensation procedure. After that, they are not used for anything at all!

The software measures low-value resistors

In order to compensate, via the **software**, the PGA103's phase shift for ranges 1 and 2 (not for range 7 and 8, as this would require high impedance with a great many stray components), we use low-value resistors (the influence of stray capacitance and/or series inductance is negligible), here too wired as closely as possible to the board, and aiming for a zero phase shift. These are only brought into use during the compensation procedure and are not used for anything else afterwards!

In this respect, adjusting the gain in this way means the voltage to be measured will change by steps in a ratio of 10. If this voltage had to be used "as-is" by the ADC, not only would the operating conditions not be ideal, but above all the measurements of the *current* and *voltage* by the ADC could be dissimilar, to the point of introducing a (severe) relative digitizing error.
To avoid this, we need to have intermediate gain values — without compromising the accuracy of the overall gain value!
Using a multiplying digital/analogue converter (DAC) (*DAC8811* in Figure 1) in conjunction with an operational amplifier (*Buffer* in Figure 1) allows us to obtain a gain that can be varied from zero to 1 (actually −1, but the phase is irrelevant here), in as many steps as the resolution of this DAC allows, and with the same precision as it!
An ingenious bit of wiring makes it possible to make the gain of this stage variable between 0 and *k* (*k* constant >1): it involves applying only a proportion of the output signal to the DAC8811's built-in feedback resistor.
If 16 gain values are used for the multiplying DAC, the basic gain variations will be selected in a ratio of $\sqrt[16]{10}$, i.e. 1.155, with the maximum gain being 1.155^{15}, i.e. 8.66. This will allow us to have a combined gain that is variable from 1 to 866 in 48 steps (see table 1).
If the gain control program is well designed, it will provide the ADC with a voltage to digitize whose maximum amplitude is close to its full-scale capacity for both the voltage and current parameters of the DUT, with a maximum amplitude difference in a ratio of 1.15.
Under these conditions, knowing that the impedance is equal to the quotient (give or take a coefficient) of the two measurements, and if the ADC has high resolution (ideally 16 bits), then digitizing errors will be practically eliminated. And there you have it as far as the gains are concerned.

Stray phase shifts

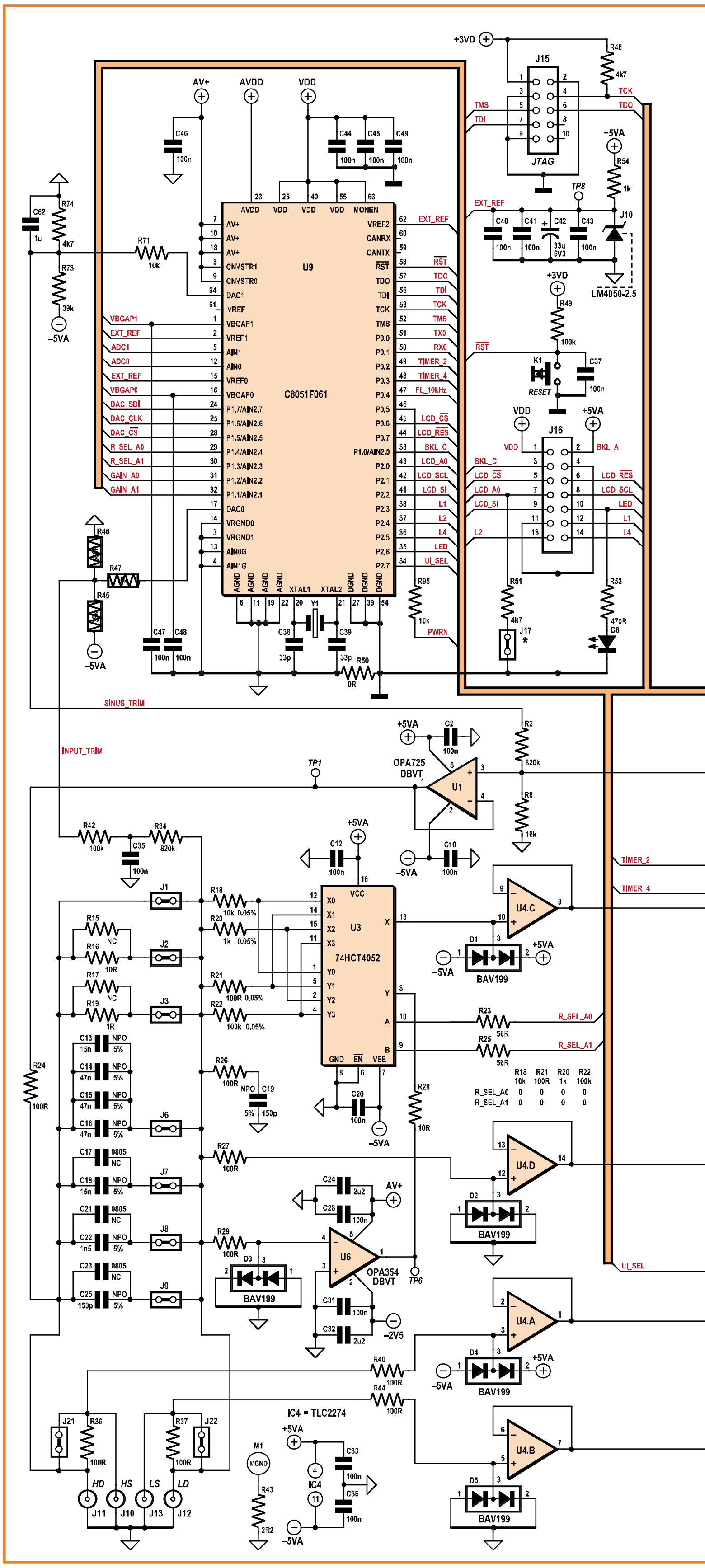
The accuracy also depends on the compensation for the stray phase shifts introduced by the measuring chain.
Two elements must be considered:

- The phase shifts introduced by the *IU converter*, as this is found only in the current measurement path. To achieve this, we're going to be using measurements of multilayer ceramic SMD capacitors with lossless dielectric (NP0 or C0G) — the compensation will be adjusted to obtain a phase shift as close as possible to the theoretical −90°. The appropriate capacitors are fitted to the PCB and brought into circuit by jumpers J6–J9.
- The differing phase shifts in the PGA103 when it is programmed for a gain of 10 or 100 in one of the measuring paths and 1 in the other: ranges 1, 2, 7, and 8 (see table below). In this case, we'll be using two SMD

resistors of 1 Ω and 10 Ω (R19 and R16), brought into circuit by J3 and J2, to achieve zero phase shift.

Measuring ranges (see table 2)
For the *voltage* and *current* measurements of the DUT impedance, eight ranges are defined from the value of R_{sense} (i.e. the *IU converter* resistor) and the gain of the main amplifier (PGA103).
The final amplification (DAC8811 multiplying DAC) between 1 and 8.66 is performed in 16 steps, denoted 0 to F, for both voltage and current measurements.

Analysing the circuit
Before looking at the circuit proper, I just want to point out that digitizing the sinewave signal directly by the microcontroller's ADC obliges us to use DC coupling from one end of the amplification chain to the other. Without such DC coupling, we'd have to wait for the ADC input to stabilize at the mean value of the signal every time the gain was changed or when switching between the *current* (*I*) and *voltage* (*V*) measurements. As the intended accuracy is ≤ 500 ppm (0.05 %), this stabilization time would be a problem. Considering the DC coupling, any offsets superimposed on the sinusoidal signal will need to be compensated so that the mean value of the signal is close to 0.000 V, even at maximum gain.
The main circuit diagram (**Figure 3**) breaks down into four sections, some of which we are already familiar with: at top left, the microcontroller; bottom left, the measuring bridge; bottom right, the amplification chain with the sinewave generator above it; and lastly, top right, the power supply and the interface with the outside world.



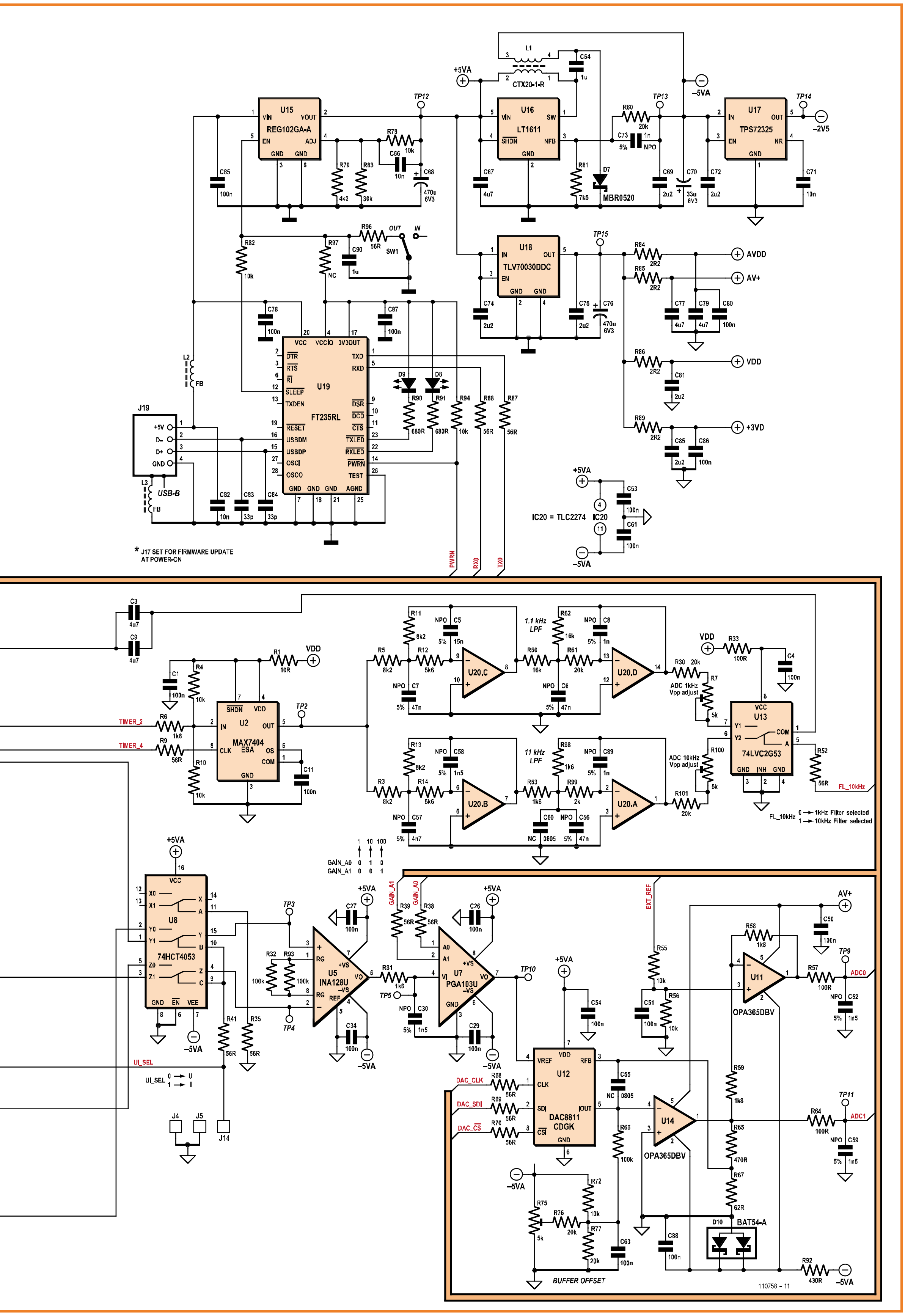


Figure 3.
Complete circuit diagram of the 500 ppm LCR Meter. Almost everything happens inside the microcontroller, but the development of the analogue part needed the greatest care in order to avoid compromising its accuracy in any way.

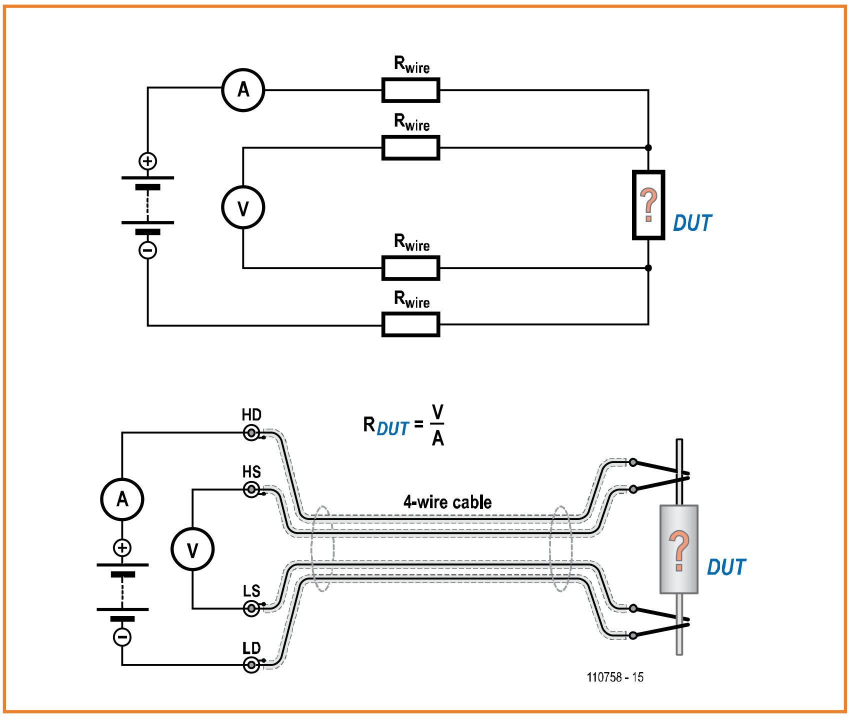


Figure 4.
The Kelvin measuring principle uses two separate wires for current flow and voltage measurement.

Measuring bridge

Having read the foregoing, you will have understood that the *IU converter* (U6, R18, R20, R21, R22) at the bottom left of the circuit diagram is a critical part of the device. The accuracy of the values of R18 and R20-R21-R22 determines the accuracy of the whole device.

The resistor R_{sense} is selected using one half of analogue switch IC U3. The buffer amplifiers U4C and D measure, at high impedance, the voltage across the R_{sense} resistor selected:

- U4D for the voltage at the junction of the resistors;
- And U4C, via the second half of U3, at the other end of these resistors (allows us to compensate for the R_{on} resistance of the switches in U3).

Hence the differential voltage between the outputs of U4C and D is equal to the voltage across the feedback resistor selected via U3.

The voltage across the DUT is measured using a 4-wire Kelvin connection: the DUT is connected between J11 and J12, i.e. the lines High Drive and Low Drive through which flow the current, while the voltage is measured, at high impedance, via J10 and J13 (the High Sense and Low Sense lines) and buffer amplifiers U4A and B.

The internal resistance of a single cable would introduce an error in the measurement result. The Kelvin clip, with its twin insulated cord, lets us perform the voltage measurement without its being interfered with by the voltage drops caused by the current flowing in the other wires (see **Figure 4**).

The dual diodes D1–D5, with very low reverse leakage current, provide overload protection.

All this is of course prone to stray capacitance, which is going to introduce phase shifts. It's essential to choose a very wide bandwidth amplifier for the *IU converter*, so as to maintain the linear relationship between phase shift and frequency. Knowing this phase shift at the maximum operating frequency (10 kHz) will enable us to calculate it at the intermediate frequencies.

The very wide bandwidth of U6 means we have to include a network to stabilize the gain at very high frequencies; this is the role of R26 and C19. Its offset is compensated using DAC0 in the µC U9: the DAC0 output voltage is programmable between 0 and 2.5 V; this is taken to a value between about −75 mV and +75 mV via the resistor network R47, R46, and R45, which will let us inject a programmable current (via R42 and R34) into U6's inverting input so as to cancel out its output voltage.

> *To check it would be possible to use the C8051F06x, I bought the development kit (C8051F060-DK) and obtained a sample of the MAX7400 (8th-order filter). I wired up the latter on prototyping board. I wanted to measure a capacitor. But how could I do this simply, using just the kit's two analogue inputs (referenced to ground)?*
> *The solution was to use two series R_{sense} + C branches, as identical as possible, connected in parallel between the filter output and ground. One with the capacitor on the ground side, the other with the resistor on the ground side. Measuring at the junction of R_{sense} and C gave the C voltage measurement on one branch and on the other the measurement of the R_{sense} voltage – and hence, of the current flowing through C!*

Amplification chain

Up front comes the analogue switcher U8 (**Figure 2**, bottom left on right-hand page) which makes it possible to select the output voltages of either U4A and B for the *voltage* measurement (*U*), or U4C and D for the *current* measurement (*I*). After this, the amplification chain is common, thus avoiding the influence of offsets and stray phase shifts; but this does mean we must use amplifiers with stable, well-defined gain.

Automatic bridge for measuring the impedance of passive components between 1 mΩ and 1000 MΩ

Differential amplifier U5 lets us go from a floating value to a value referenced to the 0 V power rail (ground). Its gain is set at 2 by R32||R93, i.e. 50 kΩ.

The gain of the precision amplifier U7 can be selected to be 1, 10, or 100 via two control lines. Then come U12, the DAC8811 multiplying digital/analogue converter amplifier, and its fast opamp U14. Taking the feedback voltage (RFB) from the divider R65/R67 rather than directly from U14 output allows us to obtain the quoted gain of 8.66. It is programmed in serial mode using 3 command lines.

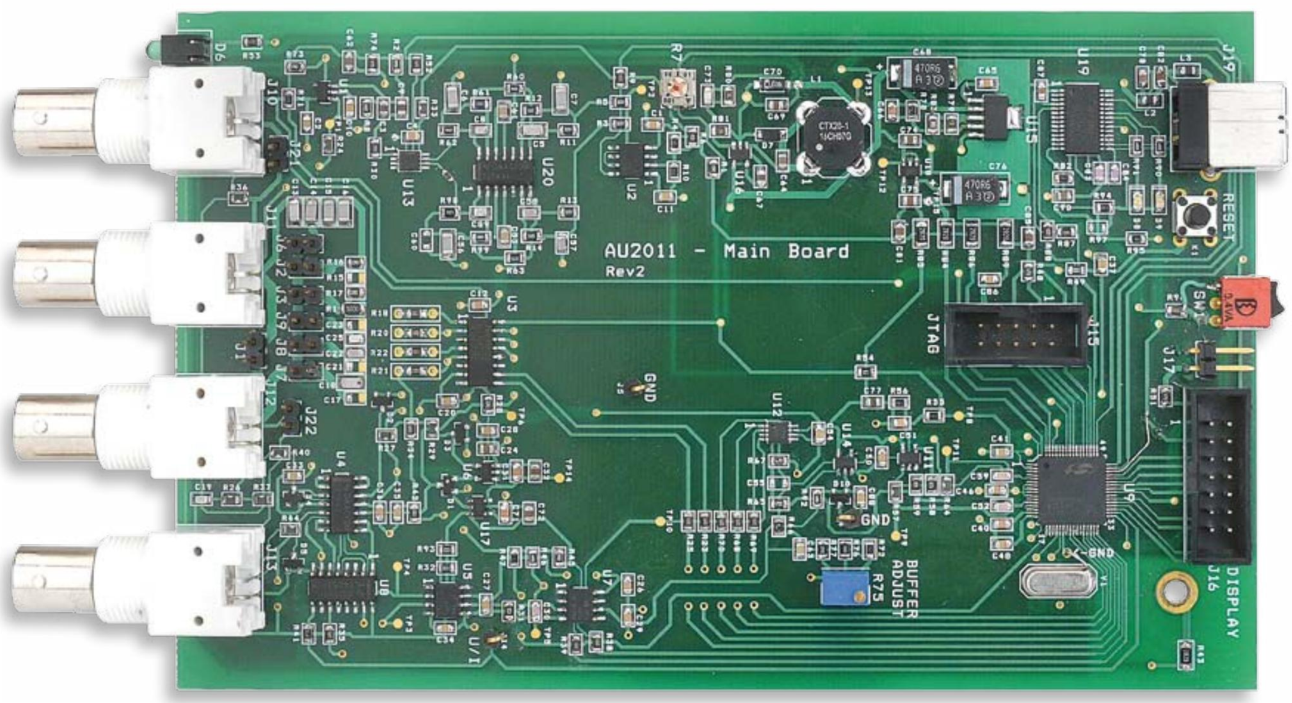
In order to make the most of their performance, the ADCs in the µC U9 are used in differential mode. To this end, amplifier U11 inverts the phase of U14's output signal.

As each of the ADCx inputs can only accept a voltage between 0 and V_{REF} (2.5 V), a DC current is extracted from the summing point so as to obtain an offset voltage of ($+V_{ref}$ / 2) and thus a differential DC voltage of 0.000 V (adjustable via preset R75, at the very bottom on the right-hand side of the circuit).

Sinewave generator

The DUT is driven by a sinewave signal that can be set between 100 Hz and 10 kHz. I use three frequencies: 100 Hz or 120 Hz (i.e. twice the AC supply frequency), 1 kHz, and 10 kHz. Other frequencies are possible (with certain restrictions). To get the most out of the microcontroller's ADCs, the digitizing process is carefully synchronized with the sinewave signal. Timer2 in the µC supplies a squarewave signal at the desired frequency, applied to the (8th order) switched-capacitor filter U2. This filter requires a clock signal (CLK) at 100× its cut-off frequency, which is provided by Timer4.

At U2's output, we have a perfectly sinusoidal signal, but we do still need to remove residual clock frequency components from it. This is the job of the two active 4th-order filters built around U20, one with a cut-off frequency of 1.1 kHz,



Operates in standalone mode with display and mini-keyboard or in USB satellite mode with a computer (OSX, Windows, Linux)

used for the frequencies ≤ 1 kHz, and the other with a cut-off frequency of 11 kHz, for the frequencies > 1 kHz. The amplitude of the 1 and 10 kHz signals is adjusted using R7 and R100. Analogue switch U13 selects which filter is used. AC coupling (C3, C9) to buffer U1 eliminates the signal's DC component.
Offset compensation for U1 is performed using DAC1 in the µC U9, in a similar way to the offset compensation for U6.

Powering

The device's supply arrives via J19 (USB-B), with +5 V on pin 1 (V_{bus}) and return on pin 4. It will come either from a USB cable connected to a computer, or from a USB power supply such as a smartphone (max. 6 V_{pp} off load).
Our device draws **more than 100 mA**. So if it is powered via a computer USB interface, this will have to be a 'high power' one, capable of supplying up to 500 mA and **guaranteeing a minimum V_{bus} voltage of 4.75 V**. It is possible to indicate the nature of our peripheral (high-power bus-powered device) by modifying the Max Bus Power parameter of the USB_Config_Descriptors in the FT245R's EEPROM, to set it to 500 mA (see the documentation for the FT_Prog utility on the FTDI website [2]).
Low-dropout linear regulator U15 supplies a voltage of +4.60 V (typ.) when its input 5 (EN) is high. Its 470 mA current limit ensures we respect the USB standard specification for a bus-powered high-power USB device.

Table 3.

$\overline{\text{PWRN}}$ State	LCD_$\overline{\text{RES}}$ State	Action
low	high	display on computer
low	low	display on computer or standalone mode if any key of the extension's mini-keyboard is pressed during boot.
high	high	ERROR! neither a computer nor the display extension is connected
high	low	standalone mode

The switching regulator U16 supplies −4.60 V (typ.). It is wired in accordance with the LT1611 data sheet [3], using a double-wound inductor L1. Regulator U18 supplies +3 V, and lastly U17, −2.5 V.
Note: On the circuit diagram, the indications +5 V and −5 V actually correspond to the +4.60 V and −4.60 V rails.

Microcontroller

The C8051F061 from Silicon Labs™ is an 8-bit analogue/digital mixed-architecture microcontroller. I chose it for the quality of its 16-bit successive-approximation ADCs ADC0 and ADC1:

- Sampling up to 1 MS/s and direct memory access (DMA);
- Max. ± 1 LSB inherent linearity in different mode;
- ± 0.5 LSB differential linearity (guaranteed monotonicity).

The rest is just as good:

- Improved-architecture 8051 core (70 % of instructions are executed in just 1 or 2 clock cycles)
- Clock frequency up to 25 MHz
- 2 12-bit DACs
- 1 10-bit, 200 kS/s ADC and 8-channel multiplexer
- 4,352 bytes of RAM and 64 KB reprogrammable flash memory
- 5 16-bit timers
- 2 serial ports (UART); data rate up to 115,200 baud (24 MHz clock)
- SMBus and SPI interfaces
- CAN 2.0 bus
- 24 general-purpose inputs/outputs
- Numerous interrupt sources
- JTAG interface, etc.

The test frequencies of 100 Hz, 120 Hz, 1 kHz, and 10 kHz are derived from the crystal-controlled 24 MHz clock.
The 2.50 V reference voltage for the A/D and D/A converters is provided by U10 (to right of the µC). Its accuracy has no effect on the overall

accuracy of the device, as its value is eliminated in the calculations.
When fitted, jumper J17 notifies the boot program of an unconditional firmware update request. LED D6 lights at the end of each valid measurement to indicate that everything's OK.

USB interface

Communication with the PC is entrusted to the well-known FT245R USB/UART converter from FTDI (U19). It's configured for 115,200 baud, 8 bits, no parity, and 1 stop bit (8 N 1) and no handshaking. LEDs D8 and D9 light when data is being transferred.
Between 25 ms and 200 ms after applying power, this IC sends its SLEEP/ pin high. If SW1 is open, U15 is then validated and all the regulated voltages are supplied to the rest of the electronics. If the device is connected to a computer via a USB cable, the PWRN/ pin, which is high when power is applied, goes low after around 300 ms. The µC's internal program tests the status of this pin during the boot sequence, along with the status of pin 6 (LCD_RES/) on extension connector J16; if this pin is low, it indicates the presence of the Display/Keyboard extension. From these two statuses, we get four possibilities (see table 3).

Temporary conclusions

This is where our initial overview of the 500 ppm LCR Meter ends. In the next part, we'll be discussing the measurement accuracy and the error or uncertainty factors (gain, calibration, phase). We'll learn about the extension with its display and keyboard, which will turn it into a standalone device, together with the computer software for the Windows, Mac, or Linux platforms), which is going to enable us to make the most of its accuracy. But you've got plenty to think about while you're waiting. And if here or there you still find some things a bit unclear, don't despair — read through it all again, check out the documentation, and things will soon become clearer.

(110758)

Links and References

[1] **Elektor** April 1997 p. 12;
May 1997, p. 12; June 1997, p. 32
All three articles can be found on the Elektor 90-99 DVD, www.elektor.com/dvd90-99.

[2] **FT245R User Guide from FTDI**
www.ftdichip.com/Support/Documents/AppNotes/AN_124_User_Guide_For_FT_PROG.pdf
[or]
http://goo.gl/USPOS

[3] **LT1611 data sheet**
http://cds.linear.com/docs/Datasheet/1611f.pdf

[4] www.elektor-magazine.com/110758

The fruits of passion

Born in 1943, Jean-Jacques Aubry, drawn at an early age to the phenomenon called Radio, developed an interest in listening-in to radio amateurs, using tube receivers. Graduating as a radio-electrical engineer in 1968, he joined a small electronics firm, where he stayed for 36 years.
He discovered computing with the ZX80 kit from Sinclair (commandeered from his young son) and taught himself programming, first with a stock management program in Turbo Pascal, then the software for various test benches in Visual BASIC. Moving on to the Mac in 1990, he applied himself to C and then C++, and discovered Qt with a Software Defined Radio (SDR).
Once retired, to keep busy, Jean-Jacques set out to design a high-performance LCR meter.

*I initially noted a discontinuity in the measured value when changing range (modifying R_{sense} or the PGA103 gain).
In the end, I realized that one of the reasons was that I'd used capacitor with X7R dielectric for C30 (filter between U5 and U7), C52, and C59 (ADC0 and ADC1 input capacitors). With the signal's alternating voltage applied, the non-linearity of this dielectric introduced noticeable distortions! The solution was to use capacitors with NP0 (C0G) dielectric.*

During my initial testing, I couldn't get the right frequencies for the timers. In the C8051F06x documentation, for the square-wave output mode, it gives (p. 298, equation 24.1):équation 24.1)

$$F_{sq} = \frac{F_{TCLK}}{2 \cdot (65535 - RCAP_n)}$$

*RCAPn being the value to be loaded into a Timer n register.
In actual fact, you need to perform the calculation using 65,536 and not 65,535!*