TIPS: Engineering Project Management
LOCATION: Canada
PAGE: 18

HOW TO: Use a Linear Sensor Array
LOCATION: United States
PAGE: 62

INSIGHT: Active Power Factor Correction
LOCATION: United States
PAGE: 80

CIRCUIT CELLAR

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SEPTEMBER 2011
ISSUE 254

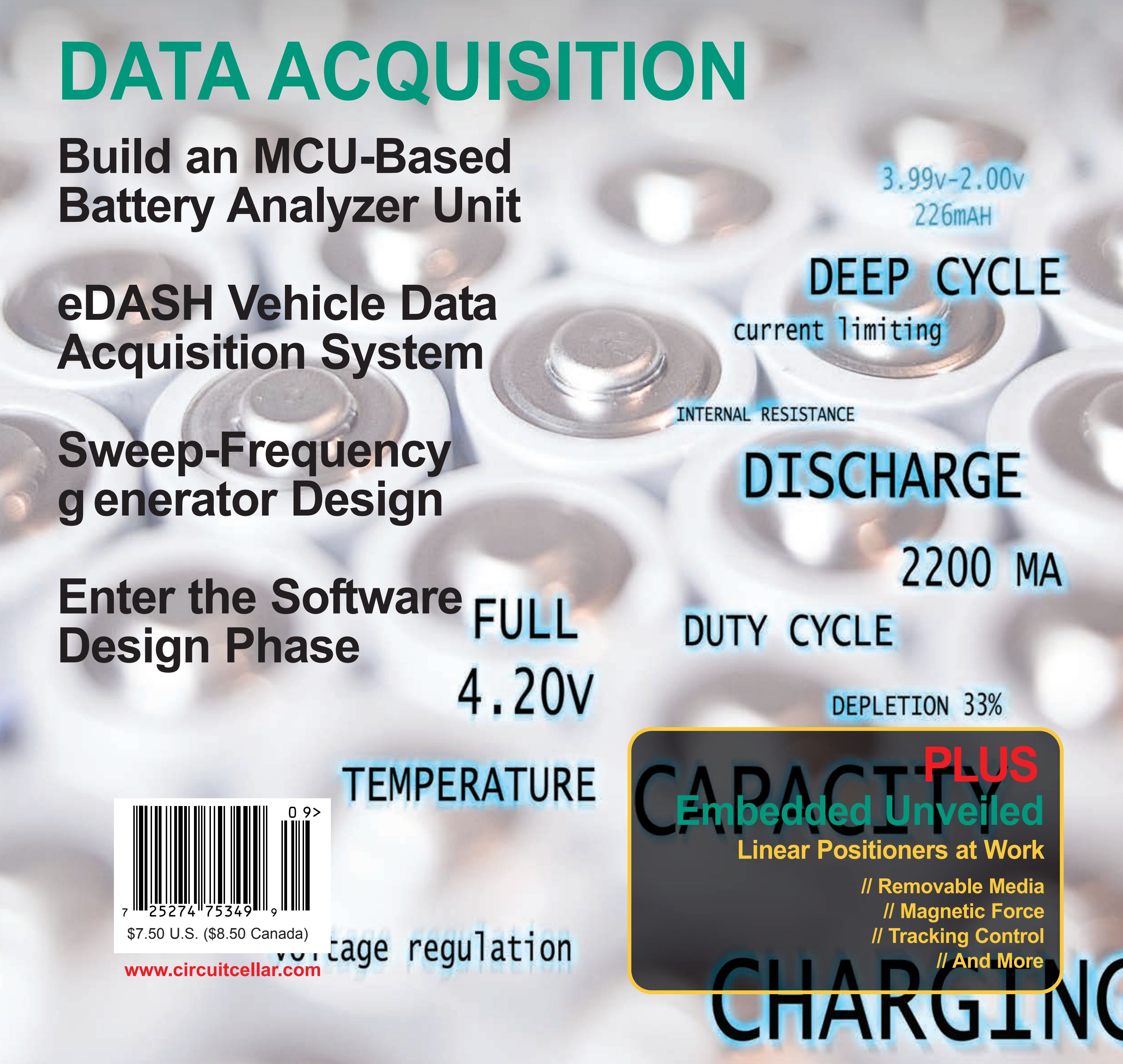
DATA ACQUISITION

Build an MCU-Based Battery Analyzer Unit

eDASH Vehicle Data Acquisition System

Sweep-Frequency g enerator Design

Enter the Software Design Phase



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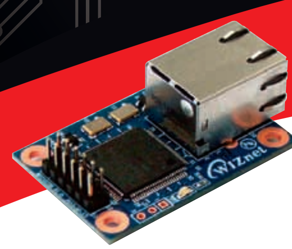
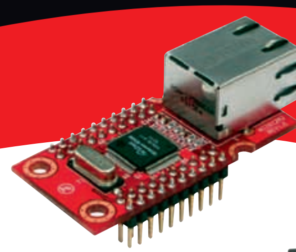
- Security System



Serial-to-Ethernet Module



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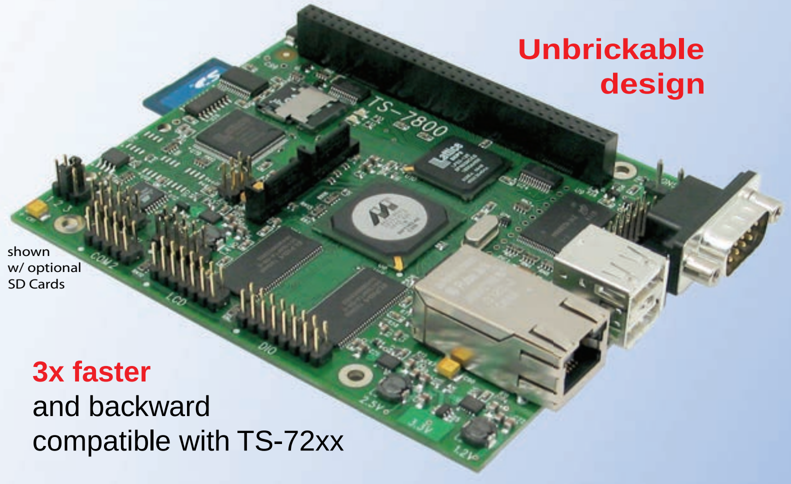

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TASK MANAGER

Inside Embedded Tech

I'm proud to announce that this issue features a new columnist. If you've been reading *Circuit Cellar* for the last few years, you're quite familiar with Richard Wotiz, who has won several international design challenges and then published insightful articles about his projects. Like Robert Lacoste before him, Richard impressed *Circuit Cellar* staffers with his innovative projects and ability to write about complex engineering in clear, concise articles.

iMCU Design

DesignStellaris Project

After Richard placed first in both the 2010 WIZnet iMCU Design Challenge and the 2010 Texas Instruments DesignStellaris Challenge, we just had to ask him to join our group of internally respected columnists.

Richard's column, titled "Embedded Unveiled," will appear every other month. In it, he'll help you investigate how electronics and embedded technologies work in real-world applications. Turn to page 56 for Richard's first column, "Linear Positioners," in which explores linear positioners and media drives in particular.

This issue's other authors are looking inside embedded technology, as well as covering various fundamental engineering topics. Let's start from the beginning.

On page 18, George Novacek highlights the importance of following proper electrical engineering practices. Both students and professionals will find his tips useful.

Testing matters. Flip to page 22 where Larry Foltzer explains how to build a digital sweep-frequency generator and a digital reference-frequency generator.

If this issue's "data acquisition" theme piques your interest, we have three excellent articles for you. One describes a battery analysis design, one is about a virtual dashboard project, and one covers how to use a linear sensor array.

Battery ratings can be inaccurate. If you frequently use li-ion batteries, implementing a design like Richard Pierce's MCU-based analyzer is a smart idea (p 34). On page 42, Robin Brophy explains how to construct the eDASH virtual dashboard, which can gather vehicle data such as speed, fuel economy, and trouble codes. On page 62, Jeff Bachiochi covers linear sensor array technology and explains how data is gathered and processed.

Lastly, we come to software design. All hardware would be useless without good software. George Martin tackles the topic of software development on page 50.

C. Abate

cj@circuitcellar.com

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Circuit Cellar's mission is to collect, select, and disseminate need-to-know information around the world in the fields of embedded hardware, embedded software, and computer applications. Circuit Cellar uses an assortment of print and electronic content-delivery platforms to reach a diverse international readership of professionals, academics, and electronics specialists who work with embedded, MCU-related technologies on a regular basis. Our aim is to help each reader become a well-rounded, multidisciplinary practitioner who can confidently bring innovative, cutting-edge engineering ideas to bear on any number of relevant tasks, problems, and technologies.

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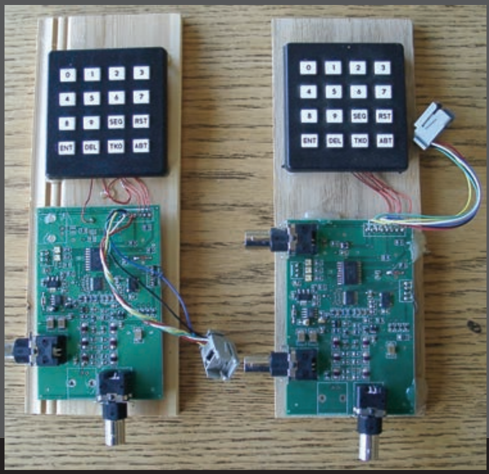
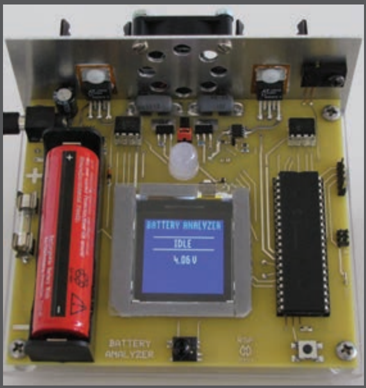
INSIDE ISSUE 254

September 2011 • Data Acquisition

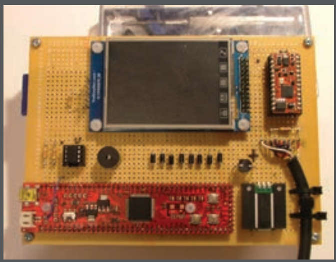
Frequency Domain Test Instruments p. 22



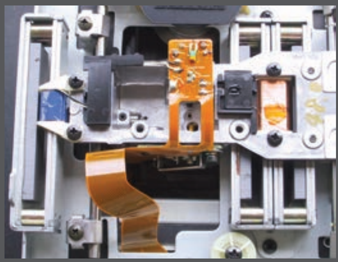
Battery Analyzer Design p. 34



Virtual Dashboard p. 42



Inside Media Drives p. 56

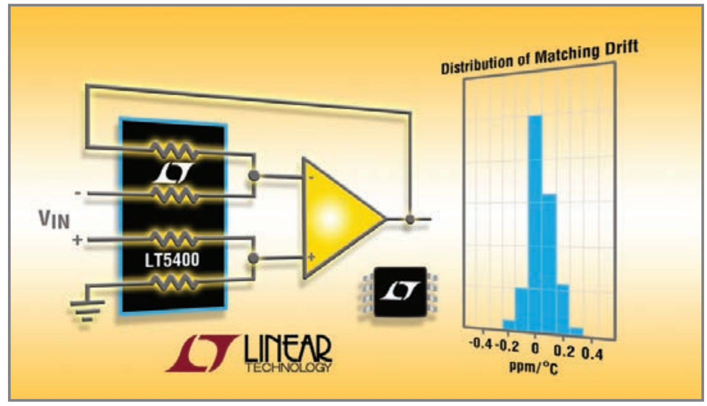


QUAD-INDEPENDENT RESISTOR NETWORKS

The **LT5400** is a family of precision matched resistors designed for high-performance signal conditioning applications in difference amplifiers, precision dividers, references, and bridge circuits. Three resistor network options are available with resistor ratios of 1:1 and 10:1, quad 10-KΩ resistors, quad 100-KΩ resistors, and dual 10-KΩ /dual 100-KΩ resistors. Each LT5400 has guaranteed matching of 0.01% from –40° to 85°C, and 0.0125% from –40° to 125°C. The LT5400 matching is enhanced with "matching for CMRR," a new metric that guarantees CMRR performance when configured in a difference amplifier circuit. The LT5400 matching for CMRR is only 0.005%.

The small LT5400 MSOP package minimizes thermal gradients and ensures consistent conditions over time and temperature for all four resistors. Since the LT5400 is based on conventional silicon fabrication and assembly, it is relatively insensitive to shock, vibration, humidity, and temperature extremes. The resistor features a high temperature option that is specified from –55° to 150°C. The LT5400 is well suited for instrumentation and test equipment in heavy industrial, military, and automotive environments.

Prices start at **$3.49** for 1,000-piece quantities.



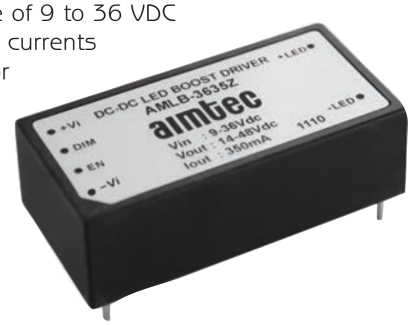
Linear Technology Corp.
www.linear.com

STEP UP LED DRIVERS

The **AMLB-Z** line of step up DC/DC LED drivers in a 2" × 1" package incorporates all the functionality and specifications required by demanding luminary applications. The AMLB-Z offers a wide (4:1) input range of 9 to 36 VDC that steps up to an output voltage range of 14 to 48 VDC and produces constant output currents ranging from 150 to 700 mA. Ripple and noise as low 350 mVpp eliminates the need for additional components.

The RoHS-compliant modular solutions feature a remote On/Off function, continuous short-circuit protection, and adjustable PWM and analog dimming functions to ensure the consistent and constant brightness of the LEDs driven.

Pricing for the AMLB-Z series of DC/DC LED drivers starts at **$13.73**.



Aimtec
www.aimtec.com

MSP430s AS LOW-COST PROGRAMMABLE METROLOGY DEVICES

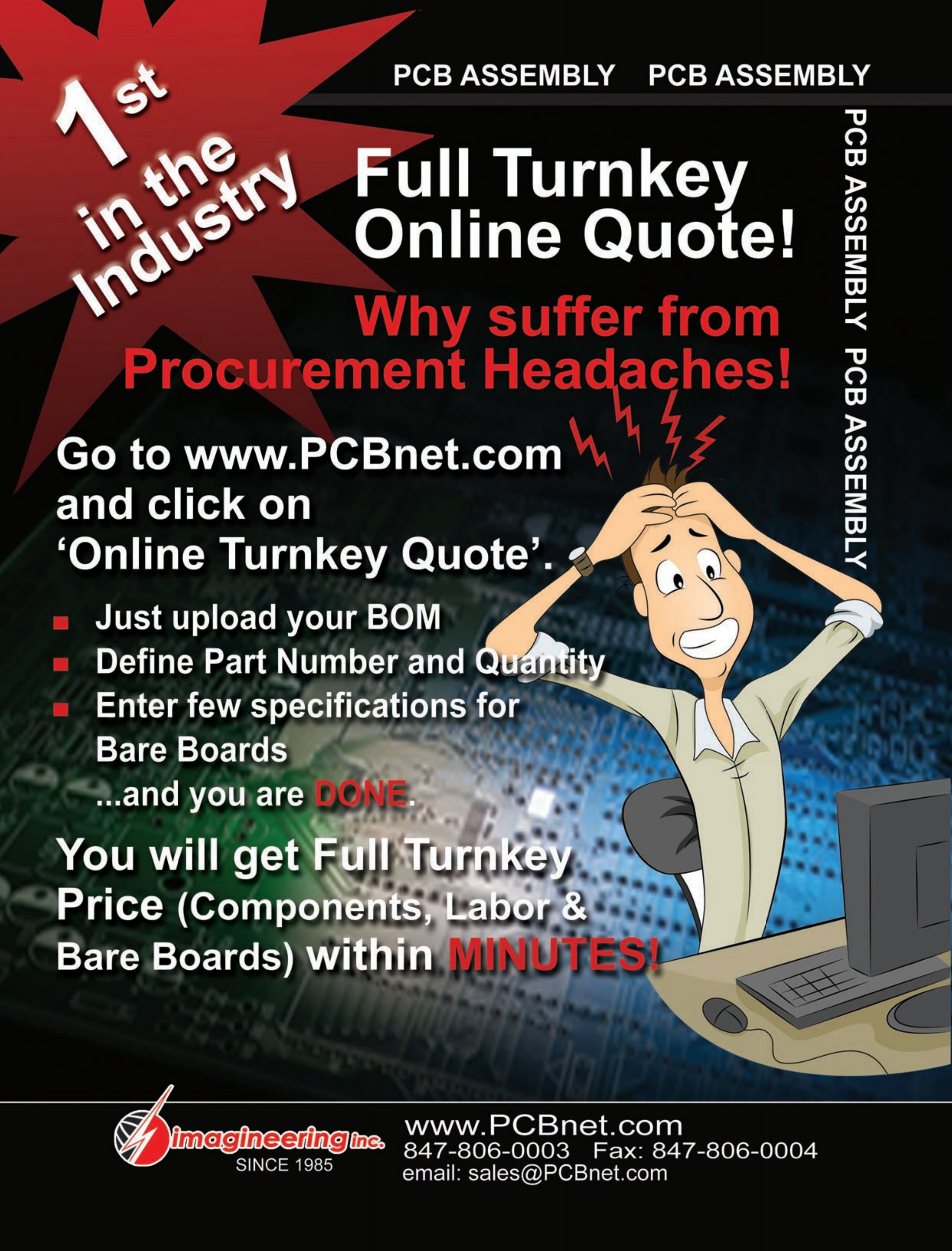
The **MSP430AFE2*xx*** is a series of metrology analog front-end (AFE) ultra-low-power 16-bit microcontrollers. The low-cost MSP430AFE series offers programmable single-phase metrology devices supported with multiple communication interfaces. The microcontrollers enable system partitioning in metering applications, such as electricity meters, home automation, submetering, and energy-saving systems, which provide flexible, standalone, high-quality measurement. The MSP430AFE series is based on a 16-bit RISC architecture with a system frequency of 12 MHz, providing high speed for increased functionality. The microcontroller achieves less than 0.1% error in energy accuracy over a wide range of 2400:1, enabled by three independent 24-bit sigma-delta converters to support antitamper.

The MSP430AFE2*xx* series is also supported by multiple tools, demos, and EVMs to provide several options for developers to begin evaluation and quickly move to production. The RF-capable MSP430 Energy Watchdog demo displays the electricity consumption of any plug-in appliance on an LCD, which enables users to manage energy usage and cost savings. The programmable MSP430AFE EVM can be used to test the new MSP430AFE2*xx* as a calibrated electricity meter. Additionally, the MSP-TS430PW24 target board and MSP-FET430U24 flash-emulation tool can be used to program and debug the MSP430AFE devices.

The MSP430AFE2*xx* microcontrollers cost **$0.80** for 1,000 units. The MSP-TS430PW24 standalone target board costs **$75**, and the MSP-FET430U24 flash-emulation tool, including software and target board, costs **$149**.

Texas Instruments, Inc.
www.ti.com

NEW PRODUCT NEWS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This series of ultra-low-power **Wi-Fi Sensors** consists of temperature, humidity, thermistor, RTD, pulse counter, and analog input modules with integrated IEEE 802.11.g radios supporting WEP 128 and WPA2 security. The wireless sensors are designed to leverage the proliferation of 802.11 infrastructures in healthcare, education, foodservice, and industrial facilities.

All products include onboard data logging in nonvolatile memory as a backup to network outages. Local audible and visual alarms alert personal of problems at the point of measurement and via the Wi-Fi network.

The sensors are designed to work with industry standard 802.11 access points for easy installation. This capability eliminates the need for repeaters, controllers, and coordinators typically needed for most wireless implementation. The 802.11 integration strategy reduces installation and ongoing maintenance costs associated with the usual overlay strategy of installing a wireless network where one already exists.

Pricing for the sensors starts at **$300**.

Point Six Wireless, LLC
www.pointsix.com

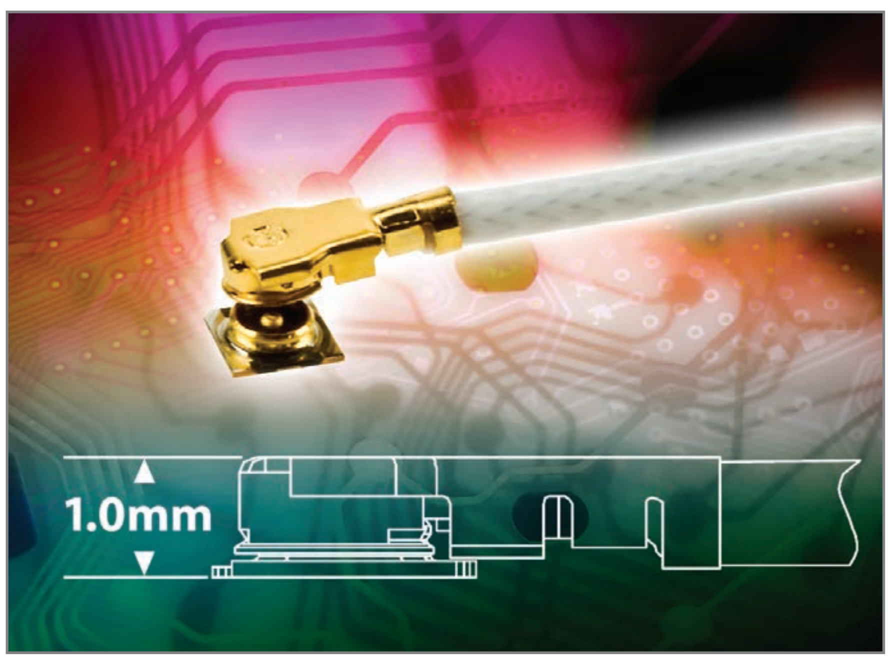
LOW-PROFILE RF MICRO COAXIAL CONNECTOR SERIES

The **JSC** series of low-profile RF micro coaxial connectors has a maximum profile of 1 mm. The connectors are designed for high-tech wireless products including tablets, smartphones, e-books, and other mobile devices. The low-profile is made possible through advanced manufacturing techniques used for both the board-mounted connector and the micro coaxial cable.

The JSC series comprises a board-mounted receptacle (MM5829-2700) that connects with a RF cable (MXJA01xxxxxx) that has a diameter of 0.81 mm. The receptacle measures 2 mm × 1.8 mm × 0.5 mm. The combined height of both the receptacle and cable plug is 1 mm, maximum. The connectors are designed to withstand up to 30 mating cycles and emit an audible click upon connection. The connector and cable structure are optimized to ensure RF performance up to 12 GHz. The RF performance has a voltage standing wave ratio (VSWR) of 1.3 GHz up to 3 GHz, 1.4 GHz maximum between 3 and 6 GHz, 1.5 GHz maximum between 6 and 9 GHz, and 1.6 GHz maximum between 9 and 12 GHz. Multiple cable configurations are available to provide flexibility and options for designers.

The JSC receptacle costs **$0.25**.

Murata Electronics North America
www.murata-northamerica.com

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The **MR330** is an absolute fiber optic position sensor (FOPS) with an all-optical design immune to any electromagnetic interference or harsh environmental conditions, such as lightening, radiation, and magnetic fields. The fiber optic aspect of the sensor also makes it well suited for long-distance position sensing over hundreds of meters without being affected by ground loop problems.

The new absolute FOPS uses an optical technique embedded in a passive sensor and an active controller connected by a duplex fiber optic link. The controller transmits a burst of light to the code disk in the sensor that accurately modulates the spectral components of the light based on angular position. The position information is imprinted in the optical spectrum of the light and guided back to the controller for precise position readout. The sensor requires no electrical power and houses no electronic components.

The MR330 measures absolute angular positioning from 0° to 360° with 13-bit (8,192 count) resolution at speeds exceeding 2,500 RPM. A complete system includes the MR332 passive sensor and the MR330 controller, which interconnect via a duplex fiber optic link. The fiber cable used is readily available industry standard 62.5/125 µm multimode fiber with duplex LC connections. The MR330-1 SSI controller offers SSI, USB, RS-485 Serial, Modbus RTU, two analog outputs (four to 20 mA and ±10 V) and two digital setpoints.

Unit pricing for a complete MR330 controller/MR332 sensor system starts at **$5,790**.

Micronor, Inc.
www.micronor.com

EXPANDED PORTFOLIO OF 5-GHZ WI-FI FRONT-END MODULES

The **RF5506** and the **RF5516** are 5-GHz Wi-Fi front-end modules (FEMs) with integrated power amplifiers (PAs). The modules deliver high power and high linearity and are optimized for the growing smartphone and tablet markets.

The high-power, high-linearity modules deliver high levels of integration in small package sizes. RFMD offers a broad portfolio of integrated FEMs that include the PA, the switch, filtering, baluns, and an optional low-noise amplifier (LNA) for both single-band (2.4 GHz or 5 GHz) and dual-band (2.4 GHz or 5 GHz) operation.

The RF5506 and RF5516 cost **$2.22** in 100-piece quantities.

RF Micro Devices, Inc.
www.rfmd.com

HIGH-DENSITY SIMULTANEOUS USB DAQ MODULE

The **DT9826** multifunction data acquisition (DAQ) module for USB combines high-performance, 24-bit analog measurement, digital I/O, counter/timers, and a tachometer channel onto a single plug-and-play device. The module is available as a board-level OEM for embedded test applications, or it can be installed in a metal BNC connection box for sensor connections.

Key design features include throughput rates of up to 41.6 ksps per channel and simultaneous operation of analog input, digital I/O, counter/timer, and tachometer subsystems at total throughput rates of more than 830 ksps. The module offers 16 simultaneous analog input channels, 16 digital I/O lines, two 32-bit counter/timer channels, one 32-bit tachometer input channel, and a 24-bit Sigma-Delta ADC per channel. The digital inputs and the two 32-bit counters can be read using the A/D subsystem and the A/D clock to synchronize digital inputs and counter timers with the analog measurements.

The unit is powered through the USB connection and offers a ±500-V galvanic isolation barrier that prevents ground loops to maximize analog signal integrity and protect computers.

Pricing for the DT9826 starts at **$1,295** for OEM quantities.

Data Translation, Inc.
www.datatranslation.com

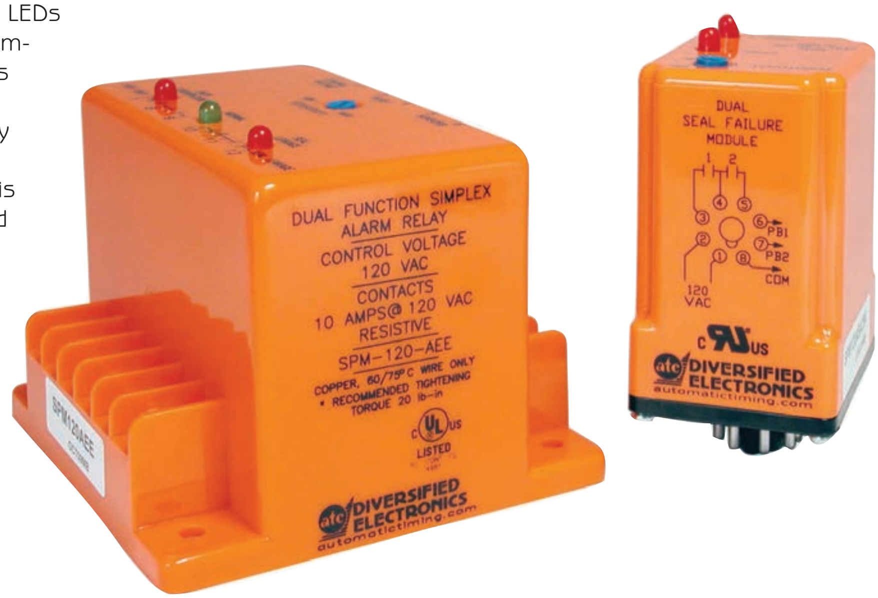
FULLY SUBMERSIBLE SHAFT SEAL & STATOR TEMPERATURE MONITOR & RELAY

The **SPM** series is a dual-function alarm and relay designed to monitor the shaft seal and stator temperature of a submersible pump motor. The series also provides early detection warning of machinery failure to reduce downtime and maintenance costs.

The SPM series is designed to operate from a 120-VAC supply voltage, with reliable operation over 10 million continuous duty cycles. The series detects leaks with a resistive float switch or a pair of conductive probes installed in the seal cavity. Over temperature is detected by a normally closed, low-temperature switch that is mounted on the stator, with LEDs that indicate "normal," "leak," and "over temperature" condition values. The SPM series also uses a bistable relay that can retain its position during power failures or energy surges. The relay automatically resets following leakage detection. Sensor voltage is 12 VDC at 50 to 60 Hz. Units are UL listed and include a 10-year comprehensive product warranty.

Pricing for the SPM series ranges from **$78** to **$138**.

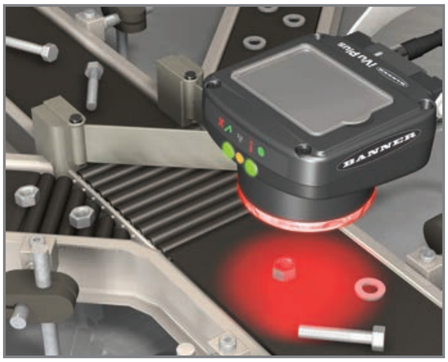
Marsh Bellofram Corp.
www.marshbellofram.com



TOUCHSCREEN VISION SENSORS DELIVER ETHERNET CONNECTIVITY

The **iVu Plus** family of touchscreen vision sensors features Ethernet connectivity. This new feature lets iVu Plus sensors share inspection data directly with PLCs, PCs, or other factory devices. The iVu Plus can store up to 30 inspections, accommodating rapid product changeover. Some models feature a Sort Sensor function that facilitates sorting up to 10 patterns in a single inspection.

Inspection setup is simple: configuration is accomplished through a menu-driven user interface, with no PC required. The iVu Plus family includes integrated touchscreen TG and barcode reader (BCR) models, as well as remote versions of both. Integrated iVu Plus sensors enable users to quickly set up and modify an inspection on-site. Remote versions pair a remote touchscreen display with one or more separate sensors, which facilitates inspection setup and monitors for areas that are difficult to access.



Each iVu Plus model is housed in an IEC IP67-rated enclosure, making the sensors rugged and versatile to suit a broad range of application environments. Cameras acquire up to 100 frames per second, and each sensor includes an integrated ring light available in red, green, blue, white, or infrared. With USB 2.0 output, users can save and load configuration data to expedite inspection setup. Additionally, with the sensor emulator users can modify an inspection offline, which reduces costly downtime.

Contact Banner Engineering for pricing.

Banner Engineering Corp.
www.bannerengineering.com

ROTARY INDUCTIVE SENSOR FOR OUTDOOR APPLICATIONS

TURCK's new line of **Rotary Inductive Analog Sensors** provides 360-degree angular measurement. Instead of using a traditional magnetic positioning element, the rotary inductive sensors use a resistance-inductive capacitance (RLC) measuring technology, which delivers complete immunity to EMC interference.

The sensors are IP67-rated, with a temperature range of –40 to 70°C. They are well suited for measuring the angle of solar panels, wind turbine blade position, crane position, and other outdoor applications. The sensors and their positioning elements are designed as separate pieces, which enables the positioning element to be mounted in a variety of unique ways (for example, directly onto a rotating shaft) without enduring the wear common to these types of sensors.

Contact TURCK for pricing.

TURCK, Inc.
www.turck.us

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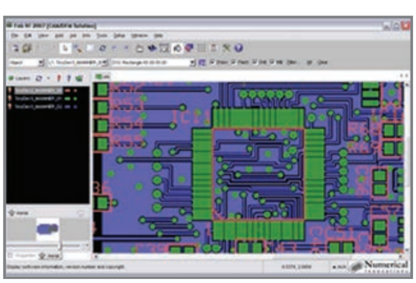


DFM Now! enables PCB designers and engineers to verify that their Gerber and drill files are ready for PCB manufacturing.

Featuring an intelligent GUI for improved speed, usability, and reliability, the software enables PCB manufacturers to instantly reach PCB designers and engineers before they send their Gerber/drill files out for manufacturing. It provides DFM Now! users with the ability to easily submit PCB designs for quotes and order placement. It also boosts CAM productivity by receiving intelligent, "ready-to-go" design data that has already passed all DFM checks and meets specific manufacturing capabilities.

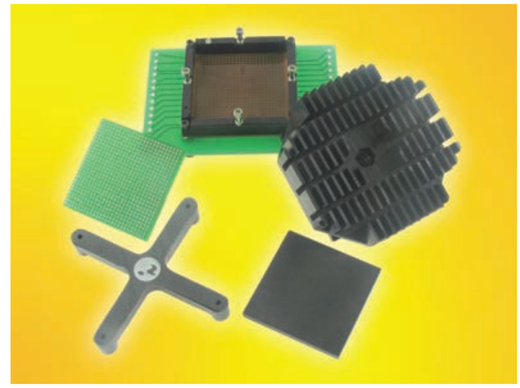
DFM Now! is completely independent and not aligned with any one manufacturer, so there is no hidden push to force designers to use a specific manufacturer. The software was designed to be an open-market system for the PCB industry, so any company can advertise inside DFM Now! free of charge.

DFM Now! is available for free download at www.dfmnow.com.

Numerical Innovations
www.numericalinnovations.com

HIGH-PERFORMANCE BGA SOCKET

The **SG-BGA-6346** socket is a high-performance BGA socket for 1-mm pitch, 616-pin BGA ICs. The socket is designed for 27 × 27 mm package size and operates at bandwidths up to 8 GHz with less than 1 dB of insertion loss. The sockets are designed to dissipate 4.5 W with passive heat sinking (natural convection) and can handle up to 40 W with active heat sinking (DC fan on top of heatsink). The contact resistance is typically 20 milliohms per pin. The socket connects all pins with 8-GHz bandwidth on all connections. The socket is mounted on the target PCB with no soldering and uses a small footprint. It is constructed with a shoulder screw and a swivel lid which incorporates a quick insertion method so that ICs can be quickly changed out.



The SG-BGA-6346 socket is constructed with high-performance and low-inductance elastomer contactor. The temperature range is –35°C to 100°C. The pin self inductance is 0.15 nH with mutual inductance of 0.025 nH. Capacitance to ground is 0.01 pF. Current capacity is 2 amps per pin. The socket works with ICs such as ST Micro FPBGA, 27 × 27 mm with 26 × 26 array and 1-mm pitch.

Pricing for the SG-BGA-6346 starts at **$807**.

Ironwood Electronics
www.ironwoodelectronics.com

LDO LINEAR VOLTAGE REGULATORS FOR AUTOMOTIVE APPLICATIONS

ON Semiconductor recently introduced five very-low and ultra-low-dropout (LDO) linear voltage regulators for a range of automotive applications, such as rear camera modules, instrument clusters, and body and chassis. These devices deliver 150 mA of output current in space-saving integrated solutions that meet the latest requirements from car manufacturers for low-quiescent currents when the ignition is switched off.

The **NCV8667** is a very-low-Iq regulator with enable, reset, and early warning. The **NCV8668** is a very-low-Iq regulator featuring window watchdog and enable and reset. The **NCV8669** is another very-low-Iq regulator with reset, delay, and early warning. The **NCV8768** is an ultra-low-Iq regulator with reset, delay, and early warning. And, the **NCV8769** ultra-low-Iq regulator features reset and early warning.

The NCV8768 and NCV8769 regulators feature typical quiescent currents of 31 µA and 25 µA respectively, making them suitable for applications permanently connected to the battery. The NCV8768 features an Enable function that further decreases the quiescent current to 1 µA, and reverse output current protection. Both devices include current limit and thermal shutdown features.

The NCV8668 features quiescent current of 38 µA and an integrated window watchdog (dual-sided detection) that enables the detection of microprocessor malfunctions. The device features quiescent current of 38 µA, delivers extremely stable performance, and requires a minimal number of external components to reduce overall solution costs and save board space. Other features of the NCV8668 include reset, a programmable delay time, and an Enable function.

The NCV8667 and NCV8669 regulators are identical, except the NCV8667 includes an Enable pin. With a user-adjustable early warning threshold option, the Iq of the devices is lowered to 28 µA typical. The devices provide replacements for ON Semiconductor's NCV4269(A) in applications where very-low-quiescent current is a requirement. Both specify a quiescent current of just 42 µA typical. With built-in reset, delay, and early warning functions, the NCV8667 and NCV8669 offer an integrated solution for a range of voltage monitoring tasks.

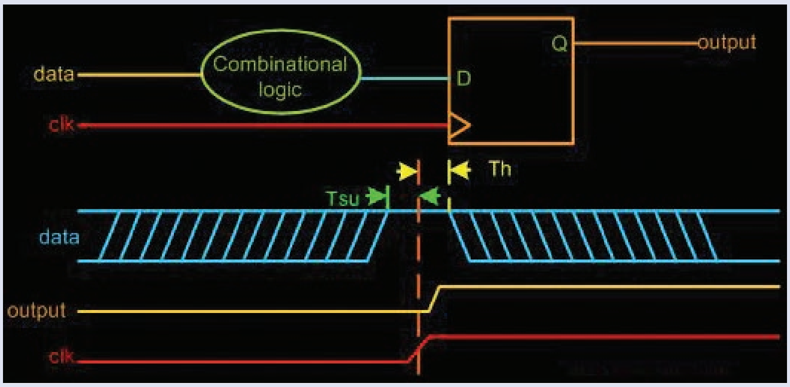
All five new parts are AEC Q100-qualified and Production Part Approval Process (PPAP) capable. NCV8768 and NCV8769 are offered in Pb-free, RoHS-compliant, SOIC-14 packages; they cost **$1.16** to **$1.32** per unit in 2,500-unit quantities. NCV8667, NCV8668, and NCV8669 are offered in Pb-free, RoHS-compliant, SOIC-8 and SOIC-14 packages; they cost **$0.82** to **$1.20** per unit in 2,500-unit quantities.

ON Semiconductor
www.onsemi.com

Test Your EQ

Edited by David Tweed

Problem 1—Why do manufacturers specify setup and hold times for flip-flops?



Problem 2—What is the fundamental principle at work in metastability?

Problem 3—Can a metastable state persist in a flip-flop for more than one clock period?

Problem 4—What does the term "plesiochronous" mean?

Contributed by David Tweed

What's your EQ?—The answers are posted at
www.circuitcellar.com/eq/
You may contact the quizmasters at eq@circuitcellar.com

by George Novacek (Canada)

Smart Project Management

Professional-level system specifications are imperative to any design project. Having project specifications in place before development saves time, cost, and effort. Here you learn why leaving things "to be determined" can lead to disaster.

In my columns, I have mentioned many times how important it is to have a thorough, well-understood system specification before starting any actual development. This time, I shall illustrate my point by demonstrating a real-life example of what can happen when this sound engineering principle is ignored. The case in point is a relatively straightforward motion control system that, because of untimely specification development, became an engineering nightmare. The system engineers, probably under the guise of concurrent engineering, issued an incomplete specification leaving many details "to be determined" (TBD) later. While this is a common practice, one detail, which ended up having a serious negative impact on the project, was left TBD.

EE CASE STUDY

Figure 1 depicts a hydro-mechanical system used to steer the nosewheel of an aircraft. It is a mechanical engineer's version of a schematic electronics engineers are familiar with. The disc shown as being rotated by the two actuators in a push/pull manner is a collar to which a strut holding the nosewheel is attached. As the collar turns, the nosewheel is steered in the desired direction. The shaded part of the diagram contains the electro-hydraulic servo valve (EHSV) we discussed last month, plus valves, pressure sensors, and other hydraulic bits and pieces we do not have to be concerned about at this time. Often, all these parts are installed in a solid chunk of aluminum with drilled channels for the fluid. Such a single part is referred to as a manifold. It is convenient to work with, but there is no reason why the hydraulic components couldn't be separately installed.

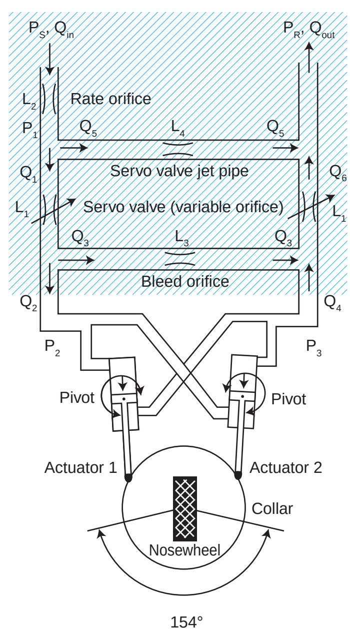


Figure 1—Example of a hydraulic steering system. Note the location of the actuator links at 120°, which is needed to maintain the required torque through the entire steering range.

Figure 1 is a classic nosewheel steering design going back to the days when the steering was manually controlled through the

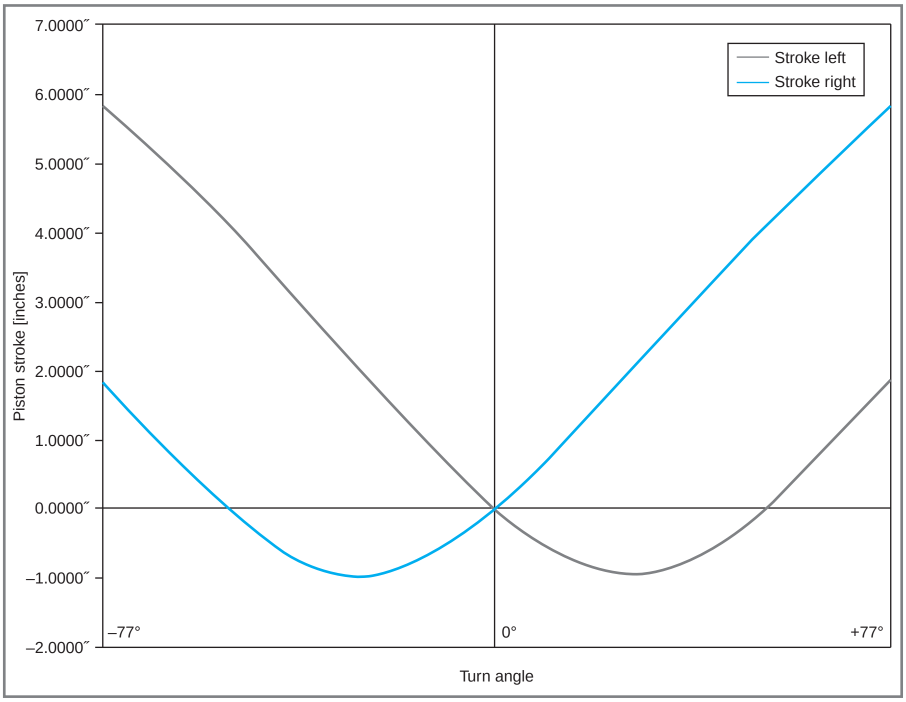


Figure 2—Piston stroke versus turn angle

engagement of control valves. Dual redundancy of the collar rotation was a fundamental requirement. Thus, there are two actuators and two mechanical links to the collar. The system engineers decided to use this time-proven design for a new, electronically controlled version and, for this reason, added two feedback position sensors. As the actuators rotating the collar pivot, the position feedback signal can be acquired by measuring the pivot angle with rotary variable differential transformers (RVDTs). Alternatively, the stroke of the piston links can be measured by linear variable differential transformers (LVDTs). Measuring the rotation of the collar itself will not work, as it defeats the redundancy requirements. More on RVDT and LVDT operation can be found in my column, "Accurate Linear Measurements Using LVTs" (*Circuit Cellar* 106, 1999).

The collar must turn ±77°, 154° in total, and a mechanical fault, such as a broken link, must be immediately detectable. In principle, it is accomplished by comparing position measurement of one actuator to the other, as they are in a defined relationship. The specification then states that the designer of the controller has to do two things. First, use the feedback data from the position sensors to generate an error signal to drive the EHSV for closed-loop control. Second, compare data from one sensor with the data acquired from the other sensor, and, if the expected values don't match, declare a fault and shut off the hydraulic power. This does not differ from countless other motion-control systems and didn't cause any concern to the electronics designers of this system either. But when the system designers eventually replaced the TBD by the feedback data relationship to the nosewheel rotation angle (see Figure 2) and the system geometry sketch (see Figure 1), we realized we had a problem.

The piston stroke versus turn angle relationship is not linear. Even worse, each side has an inversion point around –25° which gives it a useable range of about –20° to 85°. You can envision from Figure 1 how the collar needs to be pushed and pulled to achieve the desired angular turn range. The curves are mirror images of each other, but opposite in polarity. This means neither sensor can be used for steering control over the full range; both position sensors must be used. Electrical addition of the left and the right stroke signals results in an "S-curve." This is ideal for the position feedback, but its range is limited to only about ±55°. This does not satisfy the minimum required aircraft turn radius, which is a major, unchangeable characteristic for being able to turn around on a runway. Therefore, one sensor is used for steering left, and the other is for steering right. To avoid constant flipping between sensors at a straight ahead (0°) run, where most of steering occurs at high speed, the sensors change at approximately ±3°. In other words, the left sensor steers from 77° left to 3° right, where the right sensor takes over. It steers up to 77° right, then down to straight ahead and gives the control back to the left sensor at 3° left.

This is relatively simple to achieve. One caution: the position sensors must be accurately rigged to prevent jerking of the nosewheel when switching between them. The rigging offset is performed digitally, at the cost of some additional code. But, the rigging can be performed only with the aircraft lifted on jacks, when all the mechanical parts can be accurately aligned. Later drift due to wear and tear requires another rigging.

FAULT DETECTION

As I have written in the past, fault detection is often more of a challenge than the actual control. And so it is here. Imagine that you are steering left, moving on the black trace in Figure 2, which is now dominant. The black trace determines the location in the look-up table where the expected value of the blue trace is held. The retrieved look-up table value is then compared to the acquired blue trace value. If a discrepancy exists, a fault is declared. The exact opposite happens when the blue trace is dominant. Considering Figure 2, you can see that we have two problems to address. First, because of the curves' inversion, there exists an ambiguity: two steering angles show the same value on the monitoring feedback. Second, the curve around the inversion point is so flat that it does not allow for the detection of a discrepancy anywhere close to the required resolution. That means the customer's specification, based on general safety requirements, cannot be complied with. This was realized at the time when a lot of metal had been cut, a lot of money had been spent, and simply no mechanical change by the customer could even be contemplated. Numerous proposed workarounds had not been accepted for one reason or another.

In the end, the engineers were lucky. The customer, the airframe maker, and the certificating authorities accepted that at high speeds, such as during take-off or landing, the nose-wheel angle would never exceed about ±10°, including a good safety margin. Within that range, the fault detection is fully functional. Failures at low speed would not be as readily detected, but there would be plenty of time for the pilot's reaction, thus safety would not be compromised.

DISASTER PREVENTION

The cost of mechanical redesign, schedule slip, and the penalties—not to mention the loss of reputation—would have been huge. To be clear, I consider concurrent engineering to be a great method to save time and resources. But it is not a religion; there is no cookbook to adhere to. It requires management by intelligent, experienced engineers who understand the whole project and can prioritize the replacement of TBD details with data. Doing concurrent engineering just for the sake of concurrent engineering, with little understanding of all the implications, can only lead to disaster. ■

RESOURCE

G. Novacek, "Accurate Linear Measurements Using LVTs," *Circuit Cellar* 106, 1999.

George Novacek (gnovacek@nexicom.net) is a professional engineer with a degree in cybernetics and closed-loop control. Now retired, he was most recently president of a multinational manufacturer for embedded control systems for aerospace applications. George wrote 26 feature articles for Circuit Cellar *between 1999 and 2004.*

FEATURE ARTICLE

by Larry Foltzer (USA)

Sweep-Frequency Generator Design

DDS-Based Test Equipment from Sweat Equity

A proper design lab needs an accurate, programmable frequency source. And if you take a lot of repetitive frequency-domain measurements on filters, an accurate sweeper is a must-have tool. Here you learn how to use your high-speed circuit design expertise to build a digital sweep-frequency generator (SFG) and a digital reference-frequency generator (RFG).

Talk about timing! I had just completed the design and prototype implementation of two frequency-domain test instruments when I received *Circuit Cellar* 243, which featured Robert Lacoste's article, "A Tour of the Lab (Part 2): The Frequency Domain" (October 2010). Robert's article prompted me to start writing about my designs so that others might benefit from them.

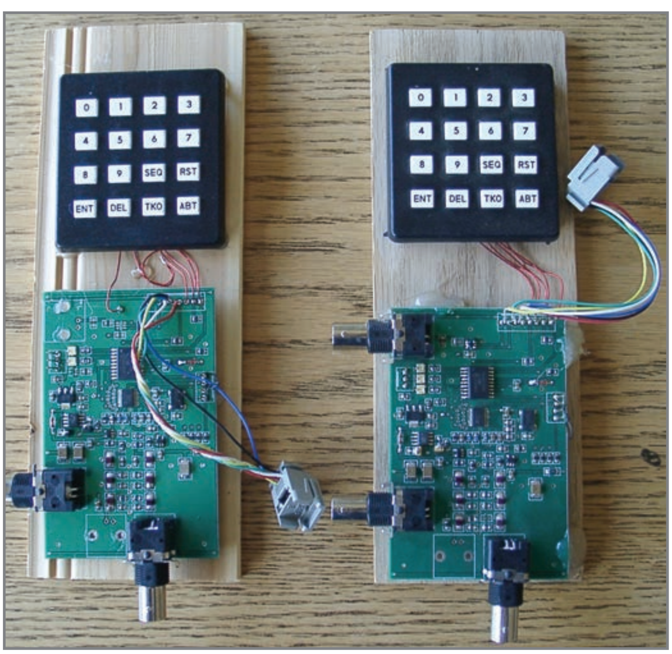


Photo 1—The completed frequency-domain test instruments. An Analog Devices AD9834-based RFG is on the left. An AD5930-based SFG is on the right. The ICSP interface used to program a Microchip Technology PIC16F627A microcontroller is provided by a dangling RJ connector socket.

As you are probably aware, Analog Devices offers a family of direct digital synthesis (DDS) devices designed to fulfill a multitude of sophisticated and important electronic functions, such as a programmable frequency reference, sweep-frequency generators, and amplitude, frequency, and phase modulators. At minimum, a well-equipped lab needs an accurate, programmable frequency source. Furthermore, if one anticipates making a lot of repetitive frequency-domain measurements on filters, an accurate sweeper is worth its weight in silver. So, armed with high-speed circuit design expertise, three pieces of silver, a Microchip Technology 8-bit microcontroller, and Assembly language programming experience, I set out to modernize and enhance the accuracy of my home laboratory with a digital sweep-frequency generator (SFG) and a digital reference-frequency generator (RFG).

From a spectral point of view, most of my home projects fall below 10 MHz. So, I selected the Analog Devices 50-MHz AD5930 waveform generator for my SFG and the 75-MHz AD9834 waveform generator for the RFG implementation (see Photo 1).

Early in the project design phase, I discovered that the signal pinout assignments of the AD9834 and the AD5930 were nearly identical, making it possible to use a single printed circuit board (PCB) design for both projects (more on that later). Therefore, I decided to start with the SFG design project since the RFG project could be viewed as a subset of the SFG from both a physical and code point of view.

The objectives of the AD5930-based SFG design were a 2- to 10-MHz frequency range, a 2-Hz frequency resolution,

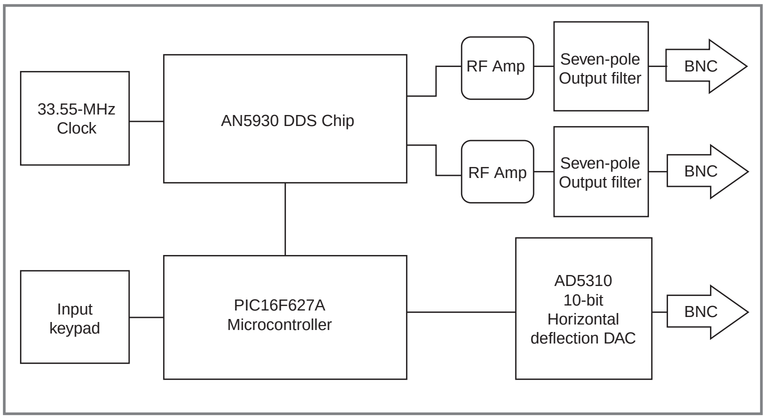


Figure 1—The functional block diagram of the sweep-frequency generator (SFG)

and an output signal level greater or equal to +7 dBm into 50 Ω. I also wanted the generator to provide a synchronized horizontal sweep signal in support of amplitude versus frequency display. The user interface had to remain simple to keep size, cost, and code complexity to a minimum.

Figure 1 shows a functional block diagram of the SFG. A description of the key features and functions follows.

KEYPAD INPUT & COMMUNICATIONS

The keypad shown in Photo 2 is a common fixture in my workshop, providing custom I/O signal functionality for numerous projects. For simplicity, the SFG uses a single LED as a handshake mechanism to signal when the system is ready for data input. When the SFG is running steady state, the handshake LED is illuminated and waits for the first key press of an input sequence. If a sweep is in progress, the LED remains lit until the current sweep is complete and the first key input is recognized. At this point, the LED is extinguished indicating that the SFG is ready for the rest of the command sequence. The LED turns on again after the appropriate parameter command key is pressed.

Data/parameter entry uses reverse-Polish notation (RPN) format. Parameter values are input in base_10 format first, followed by a character keystroke signifying what parameter is to be assigned the preceding input value. Only two input parameters are "required" to control the operation of the SFG: sweep center frequency and frequency step size. An optional third parameter may be used to increase the length of time each frequency making up the sweep (500 total) is present at the output. This duration is inversely proportional to the center frequency (F_C) and scaled/increased appropriately by the value entered here. A typical SFG programming sequence is shown in Table 1.

The white space between the first and second parameter key entry sequence shown in Table 1 is intentional. It indicates that some sweep speed and position-dependent delay may be required after the first character of the command sequence is entered in order to comply with the I/O handshake protocol. This behavior is due to the fact that the microcontroller uses inline code to control and synchronize the output frequency dwell time and horizontal output deflection signals. New input is only recognized after full sweep completion.

To set the center frequency to 4.5 MHz, enter:	`4 500000 (ENT)`
To set the frequency step to 2,000 Hz, enter:	`2 000 (DEL)`
To set the frequency step duration to 1.0 ms, enter: (This represents 200 µs overhead + 16 × 50 µs)	`16 00 (TKO)`

Table 1—A typical SFG programming sequence

The keypad scanner is based on a Microchip Technology PIC16F84 microcontroller. It is tasked with scanning the keypad for input and communicating with the SFG's microcontroller over a unidirectional serial interface consisting of the following signal and handshake lines: Data Available (`/DAV`), Serial Data, and Acknowledge (`/ACK`).

When a key is pressed, the PIC16F84 places the first bit of the keycode on the data line and pulls `/DAV` low. If the SFG's microcontroller detects `/DAV` low (`/DAV` sampled at beginning of retrace time), it reads the serial data line and pulls `/ACK` low, telling the keypad processor the data bit has been captured. The keypad then raises `/DAV` and the process is repeated until all the keycode bits have been communicated.

DDS CLOCK CONSIDERATIONS

The AD5930 is built around a 24-bit core. According to the datasheet, it can be clocked at rates up to 50 MHz. The 24-bit core size means the DDS machinery of the AD5930 can break a single sine wave into (2^{24}) discrete "phase-slices," where each phase-slice's amplitude is quantized into a 10-bit word stored in a look-up table. Sinusoidal frequency synthesis takes place by reading successive values from the phase-to-amplitude look-up table and feeding them to a high-speed DAC. The step size (SS) between successive values read from the look-up table is proportional to the desired output frequency. The following equation shows how to calculate SS. The magnitude of the phase step needed to produce F_{OUT}:

$$SS = \left(2^N\right) \times \left(\frac{F_{OUT}}{F_{DDS}}\right) \text{ for } N = 24 \qquad [1]$$

Rearranging this equation forms a ratio of constants on the left side of the equation, and a constant ratio of variables on the right side of the equation:

$$\frac{F_{DDS}}{\left(2^N\right)} = \left(\frac{F_{OUT}}{SS}\right) \qquad [2]$$

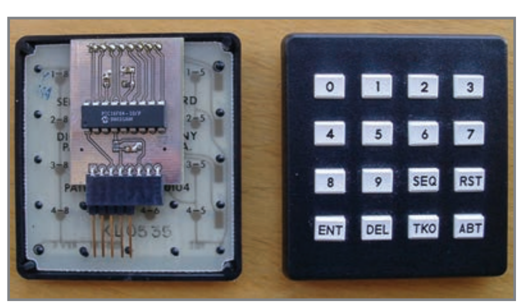


Photo 2—The keypad which features a Microchip Technology PIC16F84 microcontroller

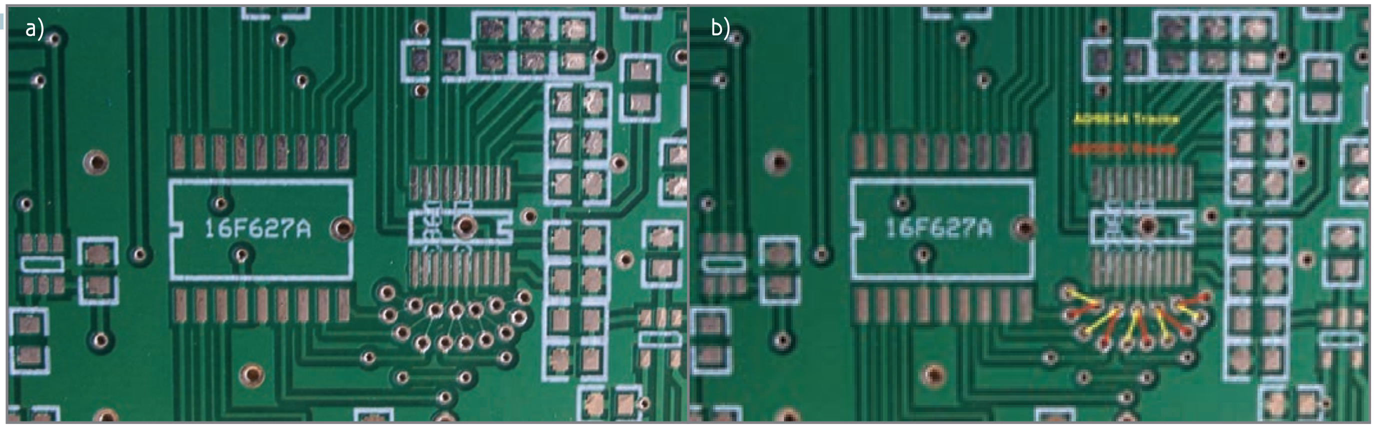


Photo 3—Magnified views of the printed circuit board (PCB). a—A virgin PCB with both sets of configuration traces intact. b—The two sets of traces used to configure the PCB are color coded to show the different traces.

For a ratio of two, $F_{DDS} = 2 \times 2^{24} = 2^{25} = 33.554432$ MHz, which is both greater than twice the Nyquist frequency and below the maximum specified operating frequency of 50 MHz. Furthermore, SS, which is the frequency programming word, must be half the desired output frequency for the ratio of two to be preserved:

$$SS = F_{OUT} \times \left[\frac{(2^{24})}{(2^{25})}\right] = \frac{F_{OUT}}{2} \qquad [3]$$

It is important to note that by employing this strategy, all you have to do to calculate the appropriate phase SS to generate a specific output frequency is divide that frequency in half. This task couldn't be easier in a binary machine. With a cleared carry bit, simply shift the frequency number to the right one bit and you're done. This makes an Assembly language programmer very happy, indeed.

Furthermore, from the equation above, you see that for a unity change in SS, F_{OUT} increases by 2, which is the frequency resolution of the system. The question is, from where do you get a 33.554432-MHz clock to drive the AD5930? The answer is that obtaining a clock source at virtually any specific frequency is easy these days, given the availability of programmable frequency references, such as the Epson SG-8002JF-PHB-ND oscillator available for just a few dollars from distributors such as Digi-Key.

DDS-TO-MCU PHYSICAL CONTROL INTERFACE

During schematic capture, I realized that Analog Devices had made an error in pinning out the programming and control signal lines on the AD5930 and AD9834. That is, the S_{DATA}, S_{CLK}, and F_{SYNC} are on different pins on these parts. Since I wanted to use a single PCB design for my two projects, I would have to provide a way to remap the necessary signals to correct this unfortunate pinout error.

Photo 3 shows magnified views of the area of the PCB where my solution is implemented. Photo 3a shows a virgin PCB with both sets of configuration traces intact. Photo 3b color codes the two sets of traces used to configure the PCB for the appropriate application. Cut the yellow set of traces for AD5930 compatibility or cut the red set of traces for AD9834 compatibility. For best results, use a number 11 surgical blade to sever the appropriate unused trace connections. If you make a mistake, through-holes are provided to facilitate a repair.

I selected a Microchip Technology PIC16F627A microcontroller to orchestrate the operation of these designs. There wasn't any particular feature that was required of these microcontrollers other than sufficient port count, low cost, and availability. And, since I already had them in my inventory, the choice was clear. You can easily port the code to other Microchip Technology parts if you are so inclined. Furthermore, you can alter the code as you see fit

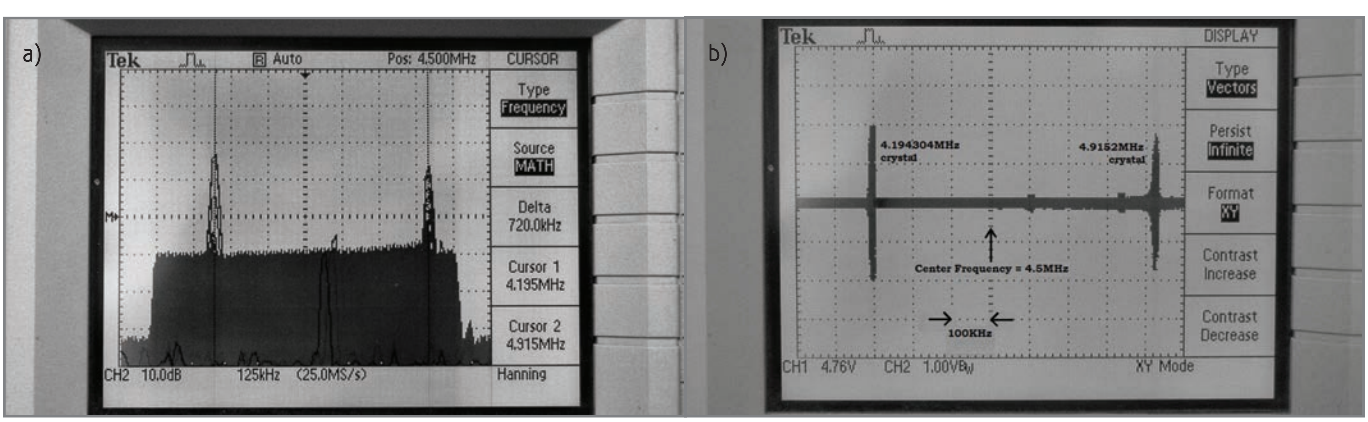


Photo 4—Sweeps configured with a width of 1.0 MHz spanning 5 to 4 MHz. The signal transmission characteristics of two parallel connected crystals connected in series with a 50-Ω terminated scope input are also shown.

and reprogram the microcontroller through the in-circuit serial programming (ICSP) interface, the dangling six-position RJ connector shown in Photo 1.

SFG DISPLAY OPTIONS

I happen to be lucky enough to own a Tektronix TDS 220, digital real-time oscilloscope that provides all the support features needed as a display for the SFG: fast Fourier transform (FFT) analysis, variable persistence, and X-Y display capability.

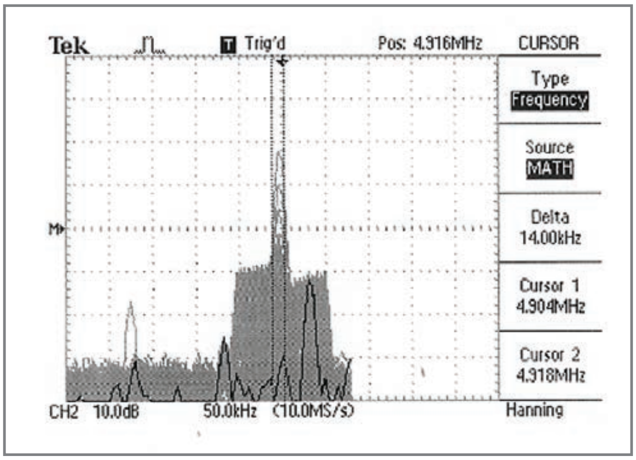


Photo 5—The crystal bandpass measuring roughly 14 kHz (about three bins wide)

In FFT mode, only a single input to the scope is required to generate an X-Y display of amplitude versus frequency. The FFT capability takes care of filling the appropriate frequency bins and displaying their contents logarithmically and sequentially along the X/frequency axis. However, while this mode offers some convenience, it suffers from poor resolution, especially in light of the spectral purity typical of a DDS source. More on this later.

In what I call the "direct approach," the TDS 220 is placed in X-Y Display mode while feeding the output of the horizontal deflection DAC to the X input channel of the scope. The sweep is always composed of 500 individual frequency steps starting at the highest frequency ($F_C + (250 \times dF)$) and ending at ($F_C - (250 \times dF)$). On an oscilloscope screen with 10 graduations, frequency display density is 50 steps per graduation. For the "`4500000 (ENT), 2000 (DEL)`" case, total sweep width is 1.0 MHz (i.e., 500 × 2,000 Hz), spanning the range 5 to 4 MHz. Photo 4 shows sweeps configured in this way, showing the signal transmission characteristics of two parallel connected crystals (4.192304 MHz and 4.9152 MHz) connected in series with a 50-Ω terminated scope input.

The difference in frequency resolution capability of the two display options is substantial. TDS 220 FFT mode resolution tracks the sampling rate (F_S) necessary to accommodate the frequencies of interest, per Nyquist. For example, to observe the transmission through a 4.9152-MHz crystal, the greatest frequency resolution is obtained when the sampling rate is set

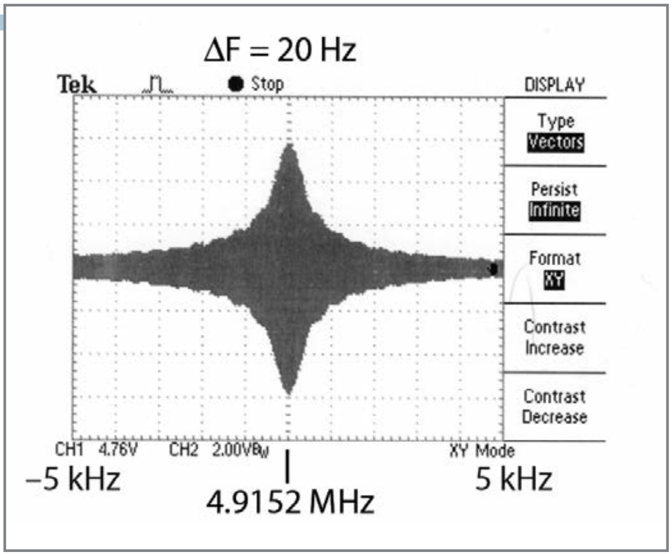


Photo 6—The transmission bandwidth measurement of the 4.9152-MHz crystal using a 20-Hz frequency step

to 10 Msps. And since the TDS 220 FFT record length is fixed at 2,048, the scope's best resolution for this signal is approximately 4.9 kHz (i.e., 10 MSps/2,048). Under these limitations, crystal bandpass measures about three bins wide or about 14 kHz (see Photo 5).

Frequency resolution using the "direct approach," is in stark contrast to the above. The direct approach, in effect, passes the signal through 1-Hz bandpass filters spaced SS Hz (effective bin width) apart. The resulting signal displays the amplitude of the RF envelope over the designated frequency span. Photo 6 shows the transmission bandwidth measurement of the 4.9152-MHz crystal using a 20-Hz frequency step (1,000 Hz per cm). Bandwidth measures roughly 30 times narrower (about 400 Hz) than obtained with the scope's built-in FFT mechanism.

The direct approach requires two inputs. The horizontal sweep deflection generator produces a linear voltage output that is decremented in sync with DDS output frequency steps. This horizontal sweep signal is generated from an Analog Devices AD5310 10-bit digital-to-analog converter (DAC). At the start of the sweep, the DAC input starts with a count of 1,012 and is decremented, two at a time, until the DAC input count reaches 12 after 500 frequency steps. Mathematically, DDS output frequency steps from $F_C + (250 \times dF)$ to $F_C - (250 \times df)$ while the sweep voltage steps from:

$$V_S = 5\text{ V} \times \left(\frac{1{,}012}{1{,}023}\right) \qquad [4]$$

to:

$$V_S = 5\text{ V} \times \left(\frac{12}{1{,}023}\right) \qquad [5]$$

Synchronization of the horizontal deflection signal DAC with the output frequency is provided by the connection of the DDS sync signal from pin 9 to microcontroller pin 11, via I/O_2 and I/O_3 shown in Figure 2.

Finally, variable persistence provides memory of the spectrum amplitude profile. This is especially useful for low-frequency measurements when the period of a single cycle could be measured in the tens of milliseconds, where the total trace

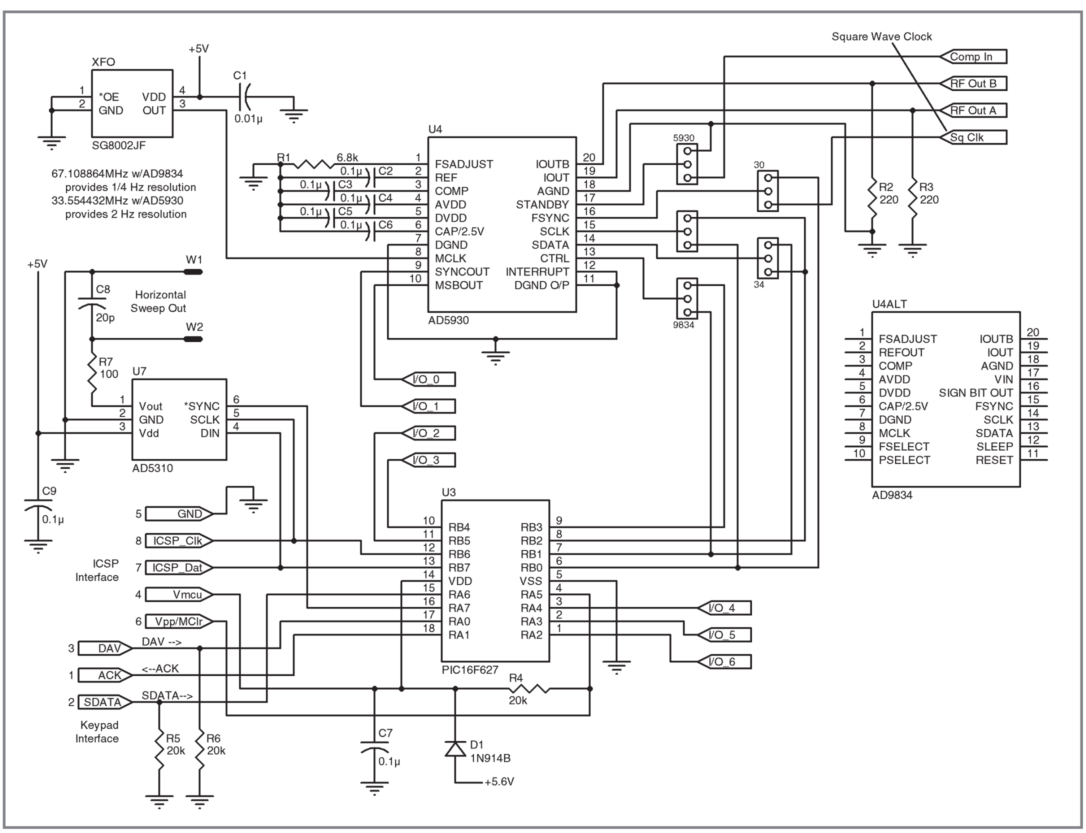
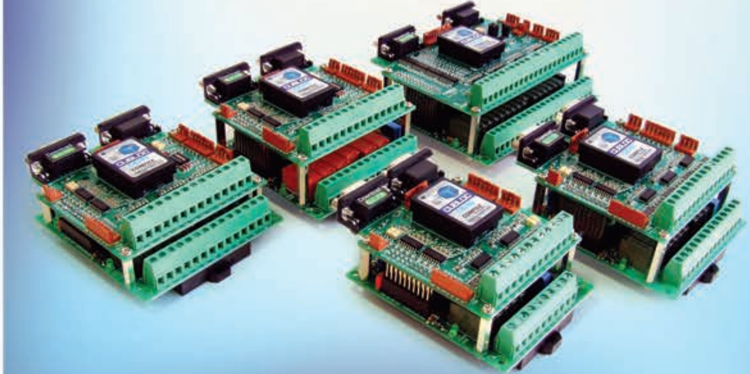


Figure 2—The synchronization of the horizontal deflection signal digital-to-analog converter (DAC) with the output frequency provided by the connection of the DDS sync signal from pin 9 to the microcontroller pin 11 via I/O_2 and I/O_3



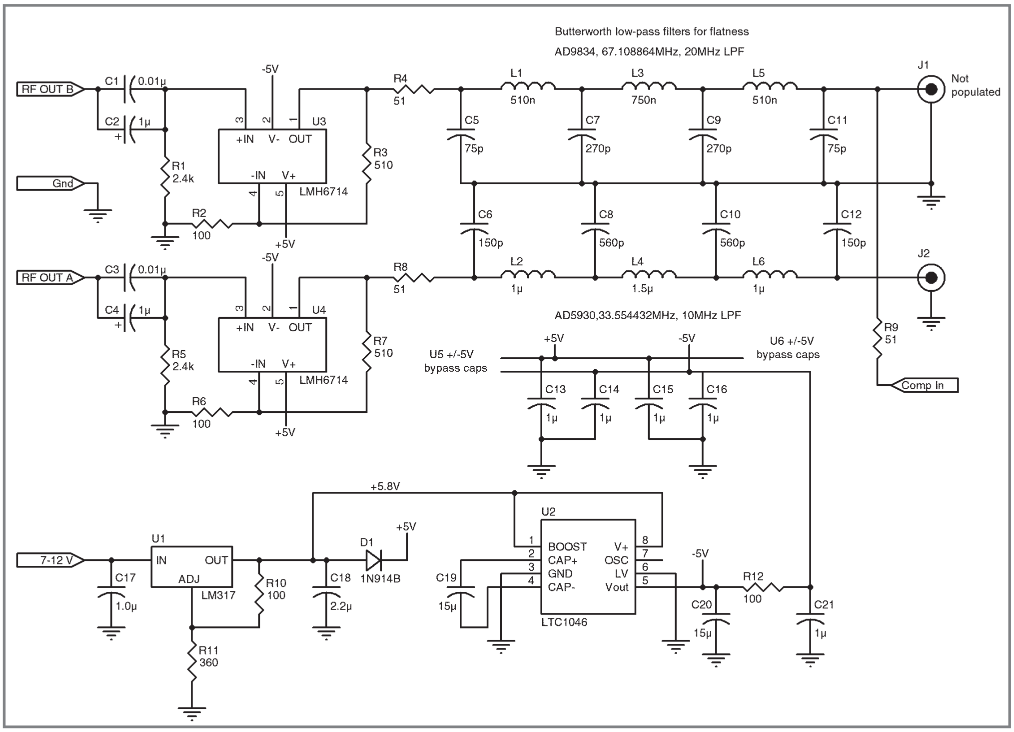


Figure 3—The operating voltages for the printed circuit board (PCB) circuits that originate from a single 7- to 9-V input via a National Semiconductor on-board LM317 regulator and a Linear Technology LT1046 converter

time for low-frequency sweeps could exceed 1 minute.

POWER CONDITIONING

Operating voltages for the PCB circuits are derived from a single 7- to 9-V input by way of a National Semiconductor on-board LM317 regulator and a Linear Technology LT1046 converter (see Figure 3). Complementary positive and negative supply voltages from the LT1046 inverter supply power to a pair of National Semiconductors LMH6714 video op-amps that subsequently drive seven-pole, low-pass output filters.

In my prototypes, I did not have the component values I had calculated for the filters shown in Figure 3, so I used the closest values I had on hand. Simulations indicated I would only pay about a 6-dB stop-band penalty for my substitutions, so I went with it. For my purposes, so far, so good!

THE INTERFACE

The SFG potentially provides two 50-Ω characteristic impedance RF outputs. In my prototype, I terminated only one RF signal in a BNC connector, but terminated the second output port with a 50-Ω resistor at the filter's output. The two RF ports are side-by-side along the bottom edge of the PCB. These outputs are capable of supplying 5 MW + 7 dBm, 1.45 V peak-to-peak into a 50-Ω load.

The BNC connector on the lower left side of the PCB, which can be turned on by software, is driven by the most-significant bit of the phase-amplitude look-up table address, producing a square wave output that is in phase with the sine wave outputs. I prefer harmonic free output, so I have it turned off.

Finally, the BNC on the upper left edge of the PCB is the output of a 10-bit DAC that generates a linear horizontal deflection signal to drive the "X" coordinate channel of an oscilloscope in sync with frequency stepping. This is a low-speed signal and intended to drive into a high input impedance. The TDS 220 enables a fine adjustment of the channel sensitivity, so that the sweep width can be adjusted to the full-scale deflection of the scope for the best display accuracy.

SFG FIRMWARE

The firmware for the SFG is available on the *Circuit Cellar* FTP site. I'm not going to elaborate on it here, except to say that I enjoy algorithm development and implementation in machine language. Even simple tasks that work well give me pleasure. Keep it simple and effective is my motto.

RFG OVERVIEW

As I stated previously, the RFG is a subset of the SFG in terms of support circuitry, firmware, and design equations. However, the faster 75-MHz AD9834 employs a 28-bit core, providing higher output frequency and frequency resolution while requiring different code to compute the frequency control word.

With the RFG based on the 75-MHz AD9834, operation to 20 MHz is possible but will require the redesign of the low-pass output filter. I did not have the appropriate parts for the 20-MHz low-pass filter design, so I used the same filter components as used on the SFG, limiting the output to 10 MHz.

The RFG and AD9834 can produce continuous wave (CW), frequency-shift keyed (FSK), and phase-shift keyed (PSK) signals with modulation externally or

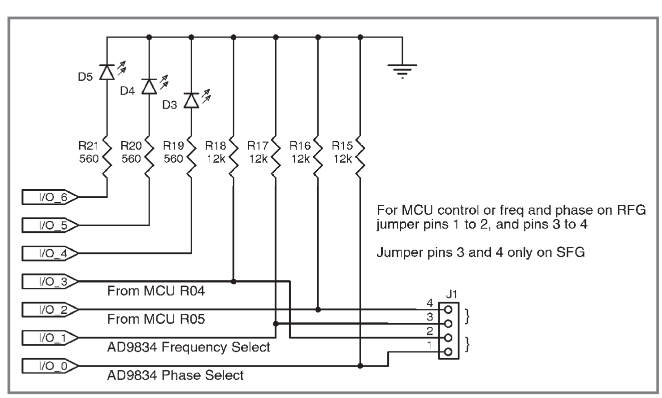


Figure 4—Jumper J1 configuration for RFG or SFG operation

by the RB4 and RB5 I/O ports on the PIC16F627A, if you write the code. Figure 4 shows how to configure J1 for modulation applications.

The output power objective is the same as for the SFG: +7-dBm output power into 50-Ω. There is no need to populate the horizontal sweep DAC, unless you are willing to write code for it for a special application (e.g., to plot the gain at a specific frequency of an AGC loop versus AGC voltage).

CLOCKING THE RFG

For convenience, you repeat Equation 2 here for N = 28, where 2^{28} is the depth of the phase-to-amplitude look-up table. Amazing!

$$\frac{F_{DDS}}{(2^{28})} = \frac{F_{DDS}}{268,435,456} = \left(\frac{F_{OUT}}{SS}\right) \qquad [6]$$

Anyway, since the maximum operating frequency of the AD9834 is 75 MHz, the resolution of the RFG, F_{OUT}/SS, will be less than one.

To maximize the output frequency range while remaining true to integer math, select 67.108864 MHz (i.e., $F_{DDS} = 2^{26}$), which sets the resolution to 0.25 Hz—that is, an incremental change of SS by four is required to shift F_{OUT} by 1 Hz. Expressed mathematically, the frequency control word is:

$$SS = F_{OUT} \times \left[\frac{(2^N)}{F_{DDS}}\right] = F_{OUT} \times \left[\frac{(2^{28})}{2^{26}}\right] = 4 \times F_{OUT} \qquad [7]$$

Now all you have to do to is the following. Calculate SS by changing the base-10 input string from the keypad to hexadecimal. Perform two PIC "RLF SS,F" instructions to multiply the desired frequency by four. Perform an "ANDLW 0xFC" instruction on the LSB to mask off bits B1 and B0, ensuring the result is ((N) – (N modulo 4)). And then program the result into F0 or F1, as appropriate.

QED!

TO MODULATE OR NOT

While the RFG and the AD9834 are capable of generating angle-modulated signals, my interest, and the current implementation, is in producing CW output for potential use as a local oscillator for radio applications. The firmware listed here configures the RFG for CW output, with static levels from microcontroller ports RB4 and RBS, driving the P_{SELECT} and F_{SELECT} lines of the AD9834. This interconnection of the microcontroller and DDS is made by connecting pin 1 to pin 2 of J1, and pin 3 to pin 4 of J1, as shown in Figure 4.

The RFG firmware contains no code to introduce phase or frequency modulation using the microcontroller. However, the AD9834 is initialized to accept externally generated TTL-level modulation input by way of I/O jumper pins I/O_0 for phase register selection and I/O_2 for frequency register selection. If you write your own modulation generator code, you can have the microcontroller drive DDS register select pin 10 (P_{SELECT}) and pin 9 (F_{SELECT}) directly and achieve modulation rates as high as 234 Kbps.

(SEG)	"Shift Key" for functional expansion
(RST)	Full RFG reset
(ENT)	Switch output to frequency specified in (F0) register
(DEL)	Switch output to frequency specified in (F1) register
(TKO)	Change output signal phase to that specified in (P0) register
(ABT)	Change output signal phase to that specified in (P1) register
(PARAM)(ENT)	Store PARAM in F0, F_{OUT} = F0
(PARAM)(DEL)	Store PARAM in F1, F_{OUT} = F1
(PARAM)(TKO)	Store PARAM in P0, Phase = P0
(PARAM)(ABT)	Store PARAM in P1, Phase = P1
(PARAM)(SEQ)(ENT)	Save PARAM in EEPROM address F0
(PARAM)(SEQ)(DEL)	Save PARAM in EEPROM address F1
(PARAM)(SEQ)(TKO)	Save PARAM in EEPROM address P0
(PARAM)(SEQ)(ABT)	Save PARAM in EEPROM address P1
Example:	Programming F0 and F1, in that order, and switching back to F0 for F0 = 4.096 MHz, F1 = 4.9152 MHz Command Sequence: 4096000 (ENT) 4915200 (DEL) (ENT)

Table 2—Alphanumeric keystroke sequences used to control the operation of the RFG

The AD9834 can be modulated by way of bit 10 (PSEL) or bit 11 (FSEL) in the DDS control register if bit 9 is cleared in the control register. However, that modulation path is slow since you have to send a 16-bit control word just to change one bit, for a maximum speed of roughly 17 Kbps.

RFG COMMANDS & CODE

Table 2 is a list of the alphanumeric keystroke sequences used to control the operation of the RFG. (PARAM) signifies a purely numeric entry representing frequency step and phase offset parameters. And, as with the SFG, RFG input command sequence is RPN-like.

The firmware/code for this project is simple and mostly static in nature. The code is available on the *Circuit Cellar* FTP site.

THAT'S A WRAP, ALMOST

The PCB is a two-sided board and was designed using ExpressPCB CAD tools. For the most part, this design worked well, although I wish I had provided large pads for some of the higher-value capacitors since 0805 versions of them are expensive and not in my inventory. I had to tombstone larger caps on small pads to avoid bankruptcy.

My original design targeted the AD5310 DAC for the horizontal displacement signal generator. But, it turned out to be in short supply, and a direct pin-compatible version was not found. I found a solution, however, in the AD5611, whose pinout matched the PCB when mounted upside down.

Finally, I have a pet peeve: What's with all these super-small packages? Isn't the PIC16F627A SOIC-18 small enough? Is the real estate savings of the DDS packages worth the handling difficulty? I don't think so. Look at the microcontroller size versus that of the DDS. Now try soldering these devices by hand. Yes, it can be done. Then again, I've been doing this stuff for a long time. But, as my eyes get older, it becomes more difficult to play in the sandbox, with or without a microscope in the workshop. No wonder kids don't go into this business of ours; they are intimidated straight out by the smallness. I wish manufactures would offer hobby packaging so the new generation would have some way to get their feet wet.

Bring back the Heathkit! ■

Larry Foltzer (lefoltzer@dishmail.net) was employed for 30 years in the fiber optics communication industry. First, he cofounded Optelecom, Inc. A decade later, his focus shifted to data and telecommunications, where for Motorola he participated in the IEEE-802 MAC subcommittee on Token-Passing access control methods. Later, he was the System Architect of the Raychem/Raynet Passive Optical Network System.

PROJECT FILES

To download the code, go to ftp://ftp.circuitcellar.com/pub/Circuit_Cellar/2011/254.

RESOURCE

R. Lacoste, "A Tour of the Lab (Part 2): The Frequency Domain," *Circuit Cellar* 243, 2010.

SOURCES

AD5930 and AD9834 Waveform generators and AD5310 DAC
Analog Devices, Inc. | www.analog.com

Epson SG-8002JF-PHB-ND oscillator
Digi-Key Corp. (distributor) | www.digikey.com

LT1046 Converter
Linear Technology Corp. | www.linear.com

PIC16F84 and PIC16F627A Microcontrollers
Microchip Technology, Inc. | www.microchip.com

LM317 Regulator and LMH6714 Video op-amps
National Semiconductor Corp. | www.national.com

TDS 220 Digital real-time oscilloscope
Tektronix, Inc. | www.tek.com

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***Circuit Cellar* 194, 2006**

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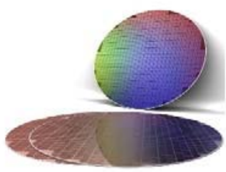

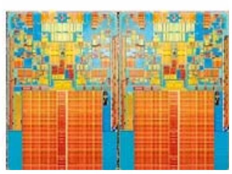

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by Richard Pierce (USA)

Battery Analysis

Build an MCU-Based Analyzer Unit

Battery ratings aren't always accurate. In fact, many batteries don't meet their rated capacities. This microcontroller-based battery analyzer enables you to sort the good from the bad. It's designed for use with single-cell li-ion batteries with a nominal 3.7-V rating, and you can use it with various cell types.

Recently, I started wondering why some of my li-ion batteries seemed to perform differently than others even though their rated capacities are the same. It's true that batteries change over time and as they are cycled, but these were all "new" batteries. So, I designed a battery analyzer to figure out what was going on.

Batteries are usually rated in mAh (or mA-h, Ah, A-h, or something similar). A mAh is one milliamp hour, and an A-h is one amp hour. A battery rated at 1 A-h (or the equivalent 1,000 mAh) means that the battery should be able to deliver the equivalent of 1 A for a period of 1 hour. For example, if the battery drain is only a 0.5 A, it should last 2 hours. If the drain is 2 A, it should last a half an hour. During the battery drain time the battery voltage is not constant; it drops from a fully charged voltage to a minimum voltage that is specified by the battery manufacturer. The discharge rate (load current) for the battery capacity rating is usually specified by the manufacturer. So you might have, for example, a 10-A-h battery with a 1-A discharge rate. If the battery is discharged at a higher rate than 1 A, it probably won't be able to provide the full 10-A-h energy rating.

My battery analyzer—which is based on a Microchip Technology PIC18F4525 microcontroller—is intended to be used with single-cell li-ion batteries with a nominal 3.7-V rating (see Photo 1). It can be used with various types of cells including 18650, 16340, 14500, 17500, 18500, and 17670 with capacity ratings of 500 to 3,000 mAh. The charge and discharge currents can be up to 1 A, maximum.

The process of analyzing the performance of a battery consists of charging the battery to a "full" level and then discharging the battery to an "empty" level while measuring and totalizing the amount of current produced by the battery. The full and empty levels are configurable, and the values can usually be found on the battery's datasheet. When it's done, the total milliamp hour discharged is shown. This value should be, in theory, pretty close to the battery rating. There are two ways that the battery can be discharged: one is for maximum energy and one is for maximum power. The maximum energy mode discharges the battery at a constant current until the empty voltage is reached, then reduces the current while maintaining the empty voltage until the battery isn't able to supply 50 mA of current. The maximum power mode discharges the battery at a constant current and then stops when the empty voltage is reached. The maximum power mode is intended to be representative of how a battery would be discharged in a real-life application.

It is also possible to charge (or discharge) a battery to a specific level. This enables a battery to be prepared for storage, for example. There are also some preset charge levels and the ability to set a custom charge level. The presets include standard (4.2 V), military (3.92 V), and storage (3.5 V). The standard level is the maximum energy that can be safely stuffed into a li-ion battery. The military level is a more conservative value that increases the battery life somewhat and enables the battery to be used over a wider temperature range while fully charged. The storage level is for storing a battery for a long period of time. Storing a li-ion battery with a high level of charge tends to reduce the battery's lifespan.

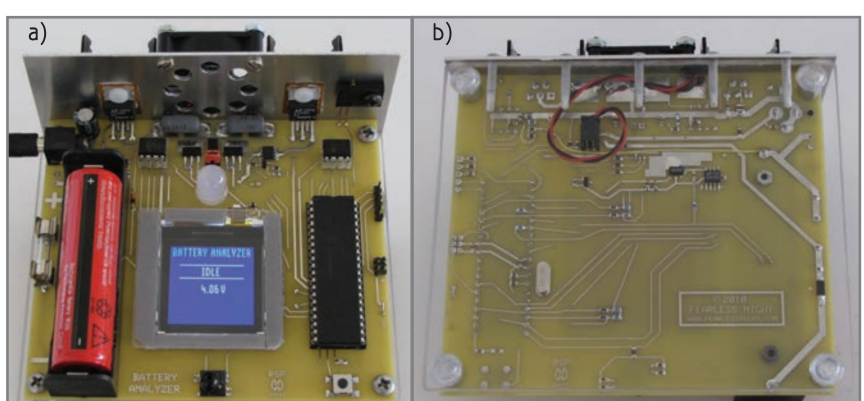

Photo 1—The front (a) and back (b) of the finished battery analyzer unit

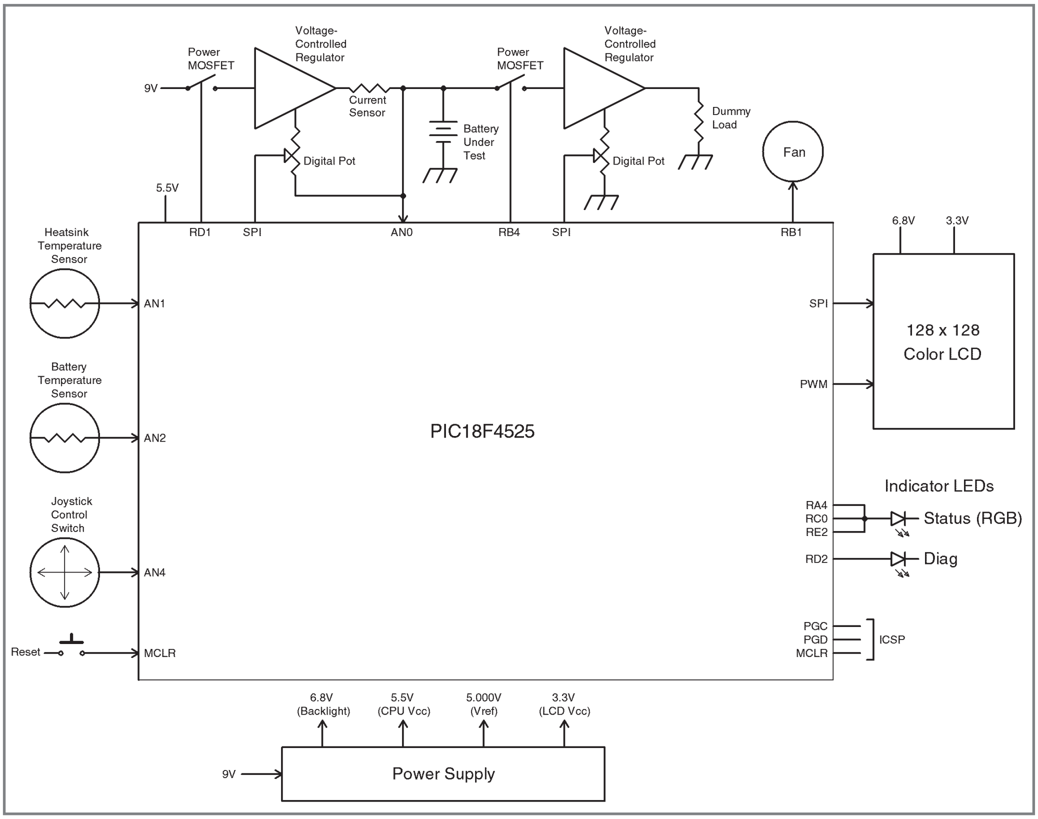

Figure 1—The battery analyzer block diagram

BATTERY CHARGING STRATEGY

A variety of li-ion battery charging methods can be used. My device uses the constant current, constant voltage (CCCV) method, where a constant current is used to charge the battery to the "full" voltage and then the charge current is reduced while maintaining the final voltage. The charging process stops when the charge current drops to 50 mA. Although it's not the fastest method, it's safe and gentle and results in a fully charged and stable battery condition.

Most li-ion batteries specify that they should be charged at a 0.2C rate (the battery rating divided by two). So, the charge rate for the battery analyzer is fixed at 0.2C. The specified discharge rate can vary; it's usually specified at something like 0.5C to get the most energy out of the battery. But, you may want to discharge at a higher rate, so this value is configurable. (Refer to the Sidebar, "Li-ion Essentials," for more information.) The maximum discharge rate is limited to 1 A. So, for example, discharging a 2,000-mAh battery at a rate above 0.2C is not possible with this battery analyzer.

POWER SUPPLY

A 9-V, 2-A AC/DC power adapter is the main power source of the battery analyzer. Good old LM317 voltage regulators are used to derive the 6.8- and 5.5-V supplies. For this application, I wanted these supplies to be pretty much right on the target voltages, and the voltages on the breadboard version were perfect. But something was different on the PCB version, perhaps due to the different LM317 package type or changing from 1/4 watt to SMT resistors, so trimming resistors were needed. The 3.3-V supply is not critical and is low-current, so a 78L33 does the job nicely. A precision 5-V reference chip provides the analog voltage reference.

Normally, a 5-V logic supply would be used, but the 5.5-V logic supply voltage level is necessary for the digital potentiometer chips so their wiper voltages don't exceed the supply voltage. The CPU and potentiometer chips are rated for this voltage, but there is no margin for error.

CPU

A Microchip Technology PIC18F4525 microcontroller is the brain of this battery analyzer (see Figure 1). I chose a crystal oscillator source so that reasonably accurate timekeeping would be possible. The LCD ended up determining the required crystal frequency. This is because the optimal speed for communication with the LCD is with a 10-MHz SPI clock. The PIC18F4525 has a four-clock instruction cycle (the instruction cycle is also the maximum SPI clock rate), an internal PLL that multiplies the 10-MHz crystal to 40 MHz (resulting in a 10-MHz SPI clock rate), and an instruction cycle rate of 10 MIPS.

The SPI is used with both the LCD and the digital potentiometers. To simplify the software, two separate SPI channels are used. The PIC18F4525 only has one hardware SPI channel so the second channel is bit-banged with software.

Li-ion Essentials

Battery capacity ratings are usually specified at a discharge rate of 0.2C (1/5 of the rated capacity). So, to get the rated capacity from a battery, it should be discharged at rate that's less than 1/5 of its rated capacity. If you are going to discharge at a higher rate, then capacity should be derated. For example, at a 2C discharge rate less than 90% of the rated capacity will be available.

Keep the battery comfortably warm. To get the rated capacity from the battery, it should be operated at 23°C or higher. At low temperatures (at or below freezing), battery performance drops off quickly, especially at high discharge currents. Be careful not to drop the battery. Physical damage, such as dents or punctures, can cause the battery to fail, sometimes in a dramatic fashion. Never short the battery terminals. Doing so can cause permanent damage to the battery, and maybe you, too.

HUMAN INTERFACE

The controls for the battery analyzer are basically two buttons. A Reset button provides a quick way to shut down everything, enter a failsafe state, and get back to the main menu. A five-way (up, left, right, down, and push) joystick button controls everything else. Although there are enough I/O pins free to directly connect the joystick, it's become a habit of mine to encode the joystick position as a voltage and then use a single analog input to read the switch state.

The display is a small (about 1″ square) color LCD. The menu-driven application program consists mainly of text displays (see Photo 2). The graphical capability is used to show the voltage and current going into (or out of) the battery over time. The display contrast is controllable by commands to the LCD chip and the backlight is controlled using the PIC PWM output and a driver MOSFET.

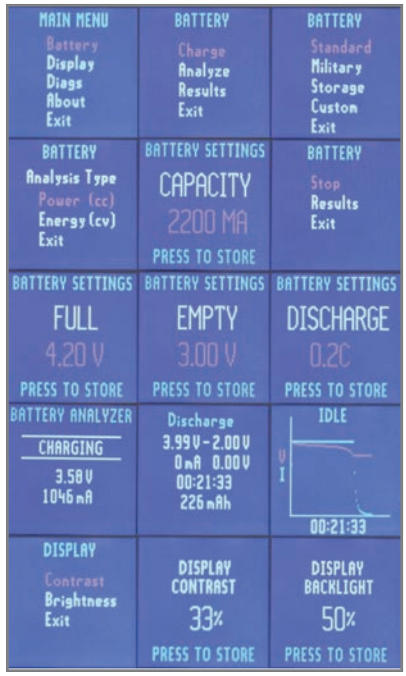

Photo 2—The battery analyzer's menu-driven display

A large 10-mm RGB status LED makes it easy to know what the battery analyzer is doing from across the room. When it's charging or discharging, the battery the LED is red; when it's done, it turns green. Blue is only used when the diagnostics are running or to indicate that a safety shutdown has occurred.

Lastly, there's the obligatory bug indicator LED (a diagnostic LED). There are only a couple of fatal errors possible in this application: a watchdog timeout or a reset instruction encounter. Should one of these errors occur, the battery analyzer will shut down to a failsafe state and then blink an error code on the diagnostic LED.

BATTERY POWER CONTROL

Two programmable power regulator circuits are attached to the battery under test (see Figure 2 and Figure 3). One circuit is for charging and the other

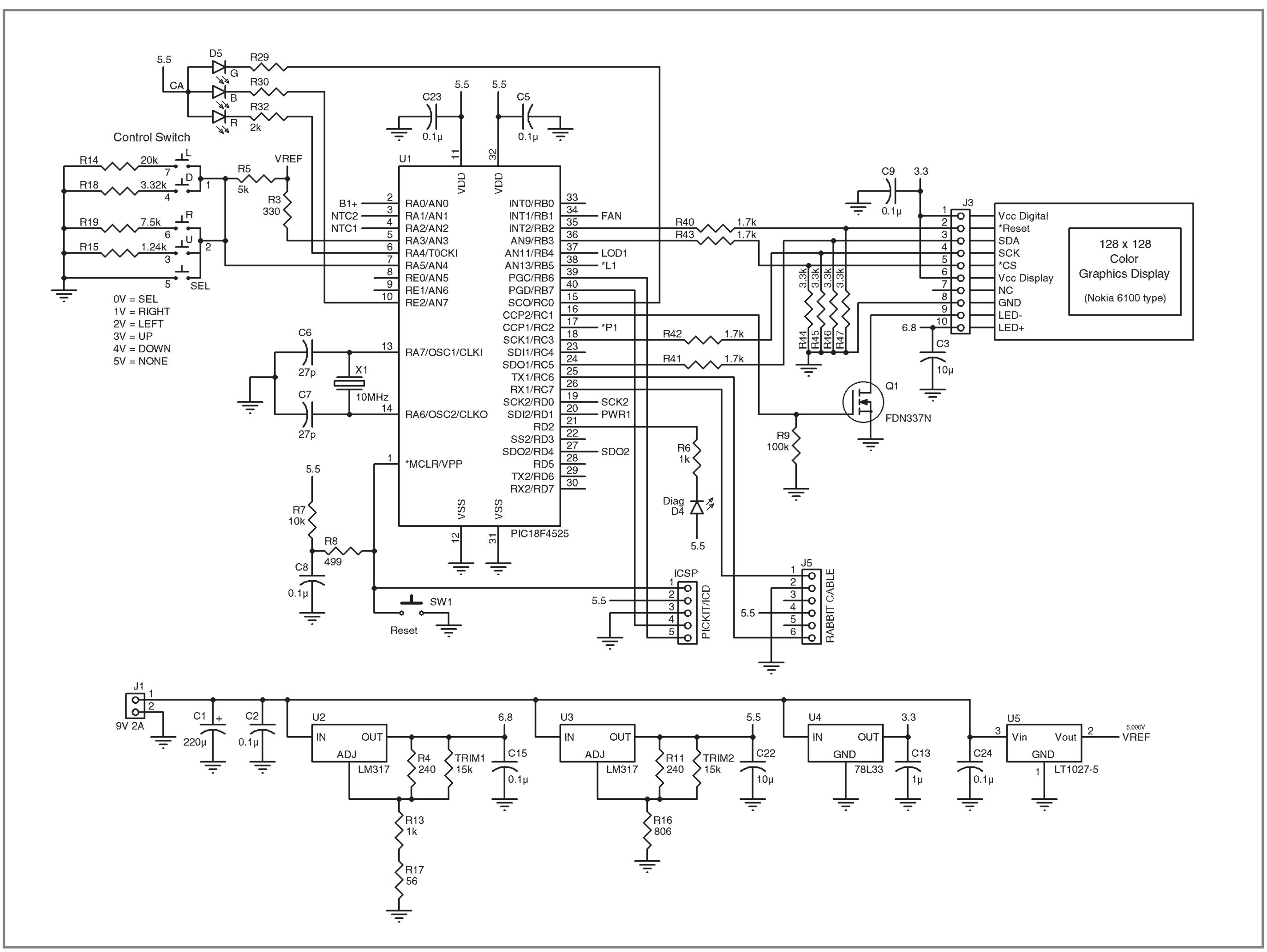

Figure 2—The battery analyzer processor, display, and power supply

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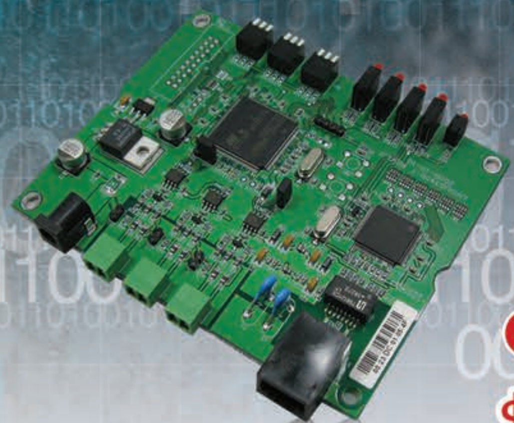

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Accessory	: Power adapter 9V 1500mA, LAN cable
Etc	: - DIP Switch(485 Baud Rate setting) - LED: Power, Network, 485 Port transmission signal

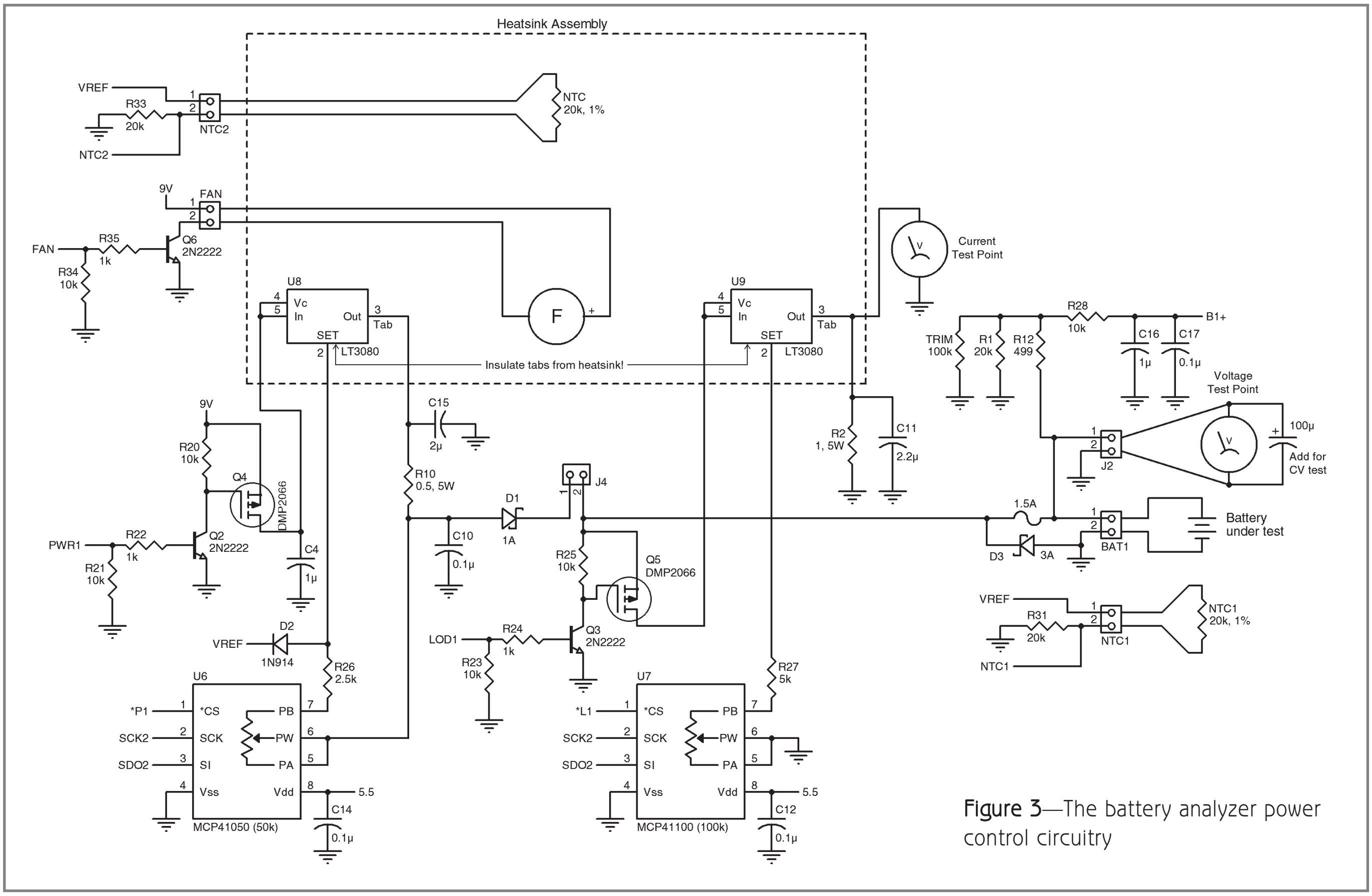


Figure 3—The battery analyzer power control circuitry

is for discharging the battery. Normally, only one of these circuits will be engaged at a time. Power MOSFETs are used to completely shut off the power regulator circuits. The MOSFETs are driven in such a way that they are normally off, so if the microcontroller is not controlling them (as during a reset) they automatically shut off.

Linear Technology LT3080 voltage regulator chips are used to regulate the battery charge and discharge currents. These voltage regulators are a little different in that the voltage Set control pin can be actively driven and the output voltage is equal to the Set pin voltage. The Set pin also can be passively used because it provides a 10-µA source current. So, to get a 5-V output, a 500-kΩ resistor from the Set pin to ground is all that's needed.

For charging, a series current sense resistor (R10) and a digital potentiometer (U6) are used to set the charge current. The 50-kΩ digital potentiometer in combination with the 10-µA source current from the LT3080 establishes a control voltage range of 0 to 0.5 V. Since the current sense resistor is 0.5 Ω, this equates to a current range of 0 to 1 A. Current-limiting resistor R26 introduces an offset of 50 mA, so the actual controllable range is 50 mA to 1.05 A.

For discharging, a constant load resistor (R2) provides a known resistance so the load current is a function of the regulator (U9) set voltage. For example, the load resistor is 1 Ω, so to get a 1-A load current the regulator set voltage is 1 V. Using a 100-kΩ digital potentiometer gives a controllable voltage set range of 0 to 1 V, which equates to 0 to 1 A. Current limiting resistor R27 introduces an offset of 50 mA, so the actual controllable range is 50 mA to 1.05 A.

Diode D1 blocks current backflow when the battery analyzer is not actively charging the battery. Diode D2 ensures that the output of regulator U8 does not exceed about 5.3 V. This is because digital potentiometer U6 could be damaged if the regulator output was allowed to go above the chip's V_{CC} supply voltage of 5.5 V.

Resistors R26 and R27 prevent excessive currents from flowing through the digital potentiometers. The LT3080 regulator, like many low-dropout regulators, has a nasty behavior characteristic at very low voltages—that is, it starts drawing a relatively large amount of current. In this case, the current would flow through the Set pin of the regulator and potentially damage the digital potentiometer chip.

Resistor divider network R1 and R12 adjusts the battery voltage to a level that's convenient for the A/D converter to use, which is 5 mV per A/D count. At least that's the theory. Actual empirical results showed a need for an additional 100-kΩ trim resistor and a couple of small filtering capacitors in order to get an accurate voltage reading. Resistor R28 prevents large currents from flowing in through the PIC18F4525's A/D input pin when the power is off but a battery is present.

THERMAL CONSIDERATIONS

I like my circuits to run cool, and in the breadboard test circuit the regulator chips were getting pretty hot. So, the thermal design of the final version was definitely overkill. Both of the LT3080 regulator chips are mounted on a 5″ × 2″ aluminum plate that acts as a heatsink. A couple of conventional TO-220 heatsinks were added. As if that wasn't enough, a small fan is also attached to the plate. The fan blows air through the plate which greatly increases its cooling capacity. The current sense and load resistors are located so that the air from the fan blows directly on them, which keeps them cool, too. As a result, it barely gets warm to the touch.

For safety, there are a couple of NTC thermistors that monitor the heatsink and battery temperatures. If either exceeds a modest temperature level, it triggers a safety shutdown.

> "To charge or analyze a battery, all you need to do is insert the battery and follow the prompts on the screen to select the desired settings."

SELF-TEST FEATURE

By simultaneously turning on both of the power regulator circuits, the battery analyzer does a little self-testing. A voltmeter attached to the voltage and/or current test points can be used to independently verify the circuit operation. There is a diagnostic menu and a number of self-test choices available to check out the charge current, discharge current, and constant voltage functions.

SOFTWARE

I generally use Assembly language with Microchip processors in the 10, 12, 16, and 18 families. So, I wrote this application in 100% Assembly.

There are only two interrupt sources: the A/D converter and the periodic timer. The high- and low-priority interrupts are used to separate the two interrupts. The periodic timer interrupt kicks off an A/D read sequence every 2.5 ms and performs human interface functions every 20 ms. An A/D read sequence is where all of the A/D inputs are read and stored into memory. Most of the A/D values are handled by the foreground process, except for the battery voltage sensor.

As soon as a new A/D reading comes in from the battery voltage sensor it's immediately processed. The constant voltage function is actually being performed by software, which means the software is part of the feedback loop. So, it's important to keep the loop response time as fast as possible to get a stable constant voltage output.

A watchdog timer interlock between the timer interrupt and the foreground main loop ensures that the watchdog will trip if either the interrupt or the foreground processes crash. The foreground main loop takes care of everything else. Although the timing is not critical, the foreground is able to do overall precision timekeeping because the timer interrupt provides accurate "ticks" for the foreground to process. These ticks are queued up so that the foreground can stall for short periods as long as it eventually catches up and processes all of the ticks in the queue.

The main foreground tasks are to run the various timers, control the fan, store the milliamp hour data-logging information, perform time and temperature safety checks, monitor for battery insertion/removal, update the LCD, and handle button inputs. By doing most of the work in the foreground, it is easier to serialize all the processes and avoid potential conflicts that can happen at the interrupt level. A couple of helper functions assist with safely accessing information that the interrupt handlers update.

The total milliamp hour discharged is an average of the current discharged during the period, which is then adjusted for a 1-h period. So, the first calculation is average = total (mA)/time (hours). To adjust for a 1-h period, the average is multiplied by the elapsed time, so mAh = average × time (hours). Combining these two equations, the time term cancels out so the resulting calculation is mAh = total (mA-s)/3,600 (s). For example, if an average of 1,000 mA was discharged over a 2-h period, the total mA-s would be 7,200,000 (1,000 mA every second for 2 hours). Dividing this by 3,600 results in a value of 2,000 mAh. Once a second, the discharge current (in mA units) is added to a total accumulator. This discharge current isn't actually measured; it is derived from the discharge current potentiometer setting. A 31/23-bit divide routine from the Microchip Technology application library is used to divide the total current accumulator by 3,600. The result is the mAh value.

The graphic data logging is performed using a dynamic compression technique. This enables data from short runs (just a few minutes) to long runs (several hours) to be recorded using the same size memory buffer. When a run is started, data is recorded at a high rate of one sample per second. If the data buffer fills up, it is compacted (half the data is discarded) and the data recording rate is reduced to one sample per 2 s. This process is repeated until the run is completed. Since the maximum run time is to the order of 8 hours and the data recording buffer is 100 entries, the lowest sample rate ends up being about one sample per five minutes.

There are a lot of display screen pages, so a data-driven subsystem is used to handle the display updates. Each display page is built using macros that specify things such as screen locations, text or variables to display, and the button handler function to use for the page. To display a new page, all that's necessary is to set the current page index number to a new value, then the display subsystem takes care of the rest. Displaying variable information is handled by a callback function for each page that enables the variable data to be set up just prior to the screen being sent to the LCD. A special callback function is provided for screens that need to draw something unique, such as the graph display in this application. Library modules provide support for the EEPROM, the LCD, the digital potentiometers, and mathematical calculations.

CONSTRUCTION & OPERATION

I constructed the final version as a naked two-layer PCB mounted on an acrylic base plate. The base plate also ties the heatsink backplate to the PCB. The battery holder is a standard 18650 type that I modified by trimming the plastic holder tabs to make it easier to remove the battery. I also cut a small hole in the back of the battery holder to accommodate the NTC temperature sensor that's located directly behind the battery. The LCD screen is held in place and protected by a small acrylic window that's anchored to the PCB with some double-sided tape.

To charge or analyze a battery, all you need to do is insert the battery and follow the prompts on the screen to select the desired settings.

HAZARDS

This battery analyzer is not well protected from reverse polarity (a battery that's inserted backwards). There should be some degree of protection provided by the fuse and reverse diode (D3), but I am too nervous to actually test it. The idea is that if the battery is inserted backwards the reverse voltage to the circuit will be minimized and the fuse may blow. But if sufficient reverse voltage makes it into the circuit, then some components will most likely be damaged.

This battery analyzer also won't prevent you from doing things to a li-ion battery that you shouldn't, such as overcharging or overdischarging the battery. So, read the manual for your li-ion battery and be careful. Not following this advice could result in damaging the battery and possibly causing it to explode. Using a "protected" battery (one with an integrated safety circuit) is a good safety measure, especially if you are not familiar with the potential hazards of li-ion batteries.

YOU GET WHAT YOU PAY FOR

Battery ratings seem pretty optimistic. I haven't found a battery yet that meets its rated capacity. And I have discovered some batteries that may be defective or perhaps even counterfeit products. Perhaps a battery rating is like the top speed of your car's speedometer. On a good day, with the wind behind you and going downhill, you might be able to get the rated amount of power out of a li-ion battery. But, from now on, I'll only be buying my batteries from reputable sources and derating the battery specifications in my battery-powered designs. ■

Richard Pierce (rsp@fearlessnight.com) is a medical product design engineer who studied Computer Engineering at the University of Illinois. He also designs home entertainment/automation gadgets.

PROJECT FILES

To download the code and bootloader files go to ftp://ftp.circuitcellar.com/pub/Circuit_Cellar/2011/254.

SOURCES

TO-220 Heatsinks
Aavid Thermalloy | www.aavidthermalloy.com

LT3080 Voltage regulator chips
Linear Technology Corp. | www.linear.com

PIC18F4525 Microcontroller
Microchip Technology, Inc. | www.microchip.com

NTC Thermistors
Murata Manufacturing Co. | www.murata.com

LM317 Voltage regulators
Texas Instruments, Inc. | www.ti.com

by Robin Brophy (USA)

eDASH

A Virtual Dashboard Using OBD-II

eDASH is a virtual car dashboard that utilizes the OBD-II port and enables you to scan many different engine functions. The dashboard can be easily constructed for OBD-II-compatible vehicles using readily available modules.

This article describes the construction of an electronic car dashboard utilizing the OBD-II port. The project makes use of readily available modules that allow for easy assembly by the DIYer. It enables you to monitor various engine parameters and view or reset any diagnostic trouble codes. I call my project eDASH (see Photo 1).

The project started nearly 20 years ago when the speedometer on my car stopped working. A self-proclaimed gearhead with a degree in electronic engineering, I get a little giddy when I can combine those two fields. So, instead of trying to fix the broken speedometer, I figured I would build a digital one. I used a Motorola MC68705 microcontroller, a three-digit seven-segment LED display, a Hall effect sensor, and two magnets strapped onto the drive shaft to create the digital speedometer. After that, I had visions of a complete digital dashboard. Remember the 1980s television show *Knight Rider* and the car, KITT? That is the kind of car I wanted. However, interfacing to all the different signals I wanted would involve installing custom transducers, a lot of analog circuitry, and a complete rewiring of the car. Needless to say, that vision didn't go very far.

Fast forward a few years, when I discovered that General Motors had an assembly line diagnostic link (ALDL) connector on most of their vehicles starting in the mid-1980s. There is certain data available via this connector.

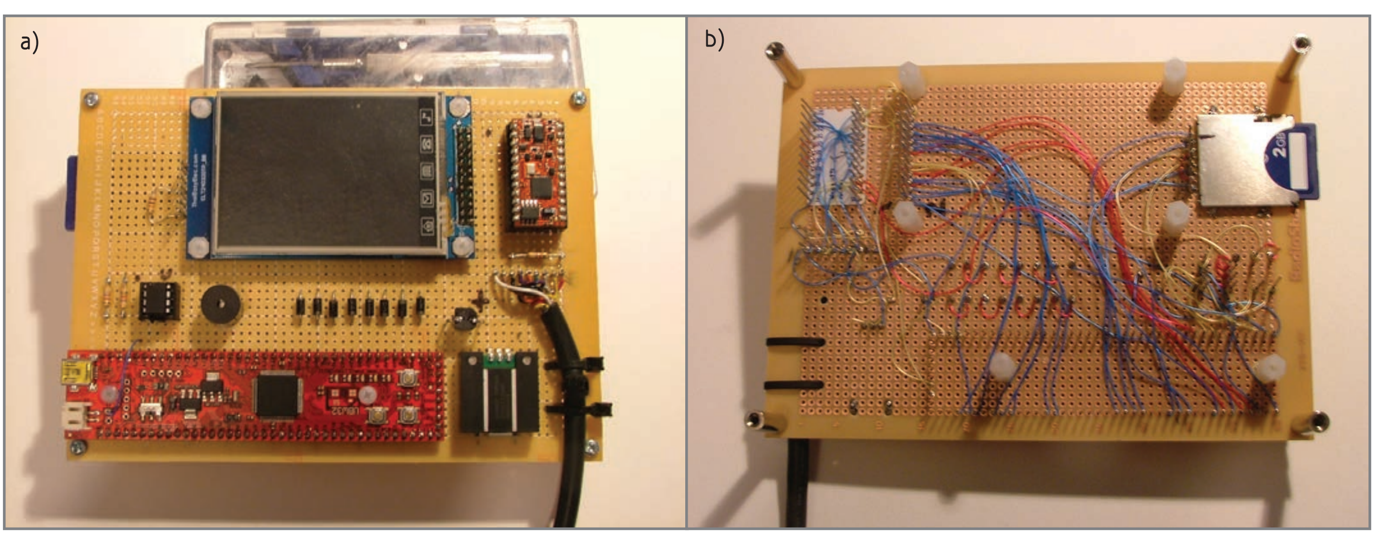


Photo 1—The top (a) and bottom (b) of the eDASH virtual dashboard

I found a software program called WinALDL and a schematic for a cable. I happened to have an '86 Camaro sitting in the garage, so I made a cable, hooked it up, and was able to scan some data. But, it was rather inconvenient having a laptop sliding around in the car. I thought about building an interface with a dedicated display, but after more research, I realized that just about every different car model had a slightly different interface. This didn't allow for a generic interface, so this idea didn't go very far, either.

Fast forward a few more years, when I discovered Elm Electronics's ELM327 OBD-to-RS-232 interpreter. OBD-II is a government-mandated standard that U.S. cars were required to comply with starting in 1996. The OBD-II standard basically states that car manufacturers must make certain operating parameters available for monitoring via a standard connector and protocol. These parameters were mainly related to emissions equipment, but basic functions such as speed, RPM, and temperature were also included. Even though there was a standard on the top layer, the lower layer was open, so each car manufacturer came up with its own interface scheme. The ELM327 interpreter took care of translating all the different schemes into a common language and essentially made every U.S. car manufactured after 1996 look like a serial port. There were commercially available interface units that utilized the ELM327 interpreter, but they all used software that ran on laptops. I learned my lesson earlier and swore I would not use a laptop. I wanted a standalone scanner with a display and input device. No computer!

I soon obtained an ELM327 interpreter and built a prototype OBD-II scanner using a Parallax BASIC Stamp microcontroller, a 7″ composite video monitor from a portable DVD player, a serial-to-NTSC video module, and a TV infrared remote. This proved that the concept of monitoring data on the OBD-II port was feasible, but I soon ran out of memory on the BASIC Stamp. The next version involved a Microchip Technology PIC18F2620 microcontroller, a Maxim Integrated Products MAX7456 single-channel monochrome on-screen display (OSD) generator, and, again, a composite monitor with a TV remote. The problems with this version were the TV remote was always getting lost and it was not easy to find a place to mount the video monitor. The next version used a bigger PIC, a 128 × 64 monochrome graphic LCD, and a small five-button pendant wired with a phone cord. On this unit, the display was limited and the pendant kept getting tangled up—you know how tangled phone cords can be, almost as bad as lost TV remotes.

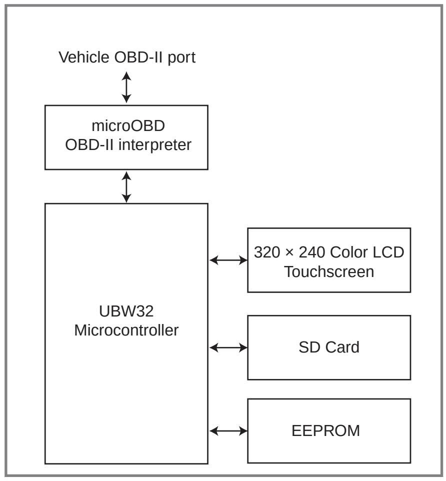


Figure 1—A block diagram of the project

The processors used in the later versions were surface-mount devices, which made it somewhat difficult to prototype. The OBD-II vehicle interface itself requires approximately 40 discrete components in order to cover all the different manufacturers' schemes. I ended up having to etch circuit boards for three different versions of the project in order to test them; there was no easy way to breadboard them.

The ultimate goal was to have something about the size of a GPS navigator with a color touchscreen. It had to be easy to build, which meant no surface-mount components, if possible. With the proliferation of MP3 players and cell phones, a supply of color LCD touchscreens is now available to amateurs and professionals alike. The last piece of the puzzle was a ScanTool.net microOBD 200, which is a complete ELM327-compatible OBD-II interpreter and interface in a standard 24-pin DIP footprint. The microOBD drastically reduces the number of discrete components required and therefore simplifies construction.

DISPLAY & DEVELOPMENT BOARD

The display module is a Gravitech 320 × 240 color LCD with a touchscreen. The module also contains a touchscreen controller as well as some buffer chips and a backlight driver. The physical connection to the display is with a 12″ × 2.1″ header. The software interface to the touch controller is via a SPI port. An 8-bit bidirectional parallel port along with several control lines connects to the display itself.

> "Using a modular approach, it is easy to construct a monitor for an OBD-II-compatible vehicle that enables you to scan many different engine functions."

I chose a SparkFun Electronics UBW32 USB 32-bit whacker development board which contains a PIC32MX795 processor, an oscillator, 3.3- and 5-V power regulators, five LEDS, three pushbuttons, and a USB bootloader. The UBW32 has all of the processor pins brought to the edge of the board for easy connection with 0.1″ headers. The PIC32MX795 is a 100-pin surface-mount device, so having the chip already mounted on a circuit board is a lifesaver. The UBW32 is basically a breakout board for the PIC. I considered some other single-board solutions, but most carry too much baggage trying to be too versatile. The UBW32 is a nice, simple board that is easy to use. A side benefit of using a PIC chip is that the MPLAB IDE and C compiler are completely free!

The microOBD module contains the OBD-II interpreter

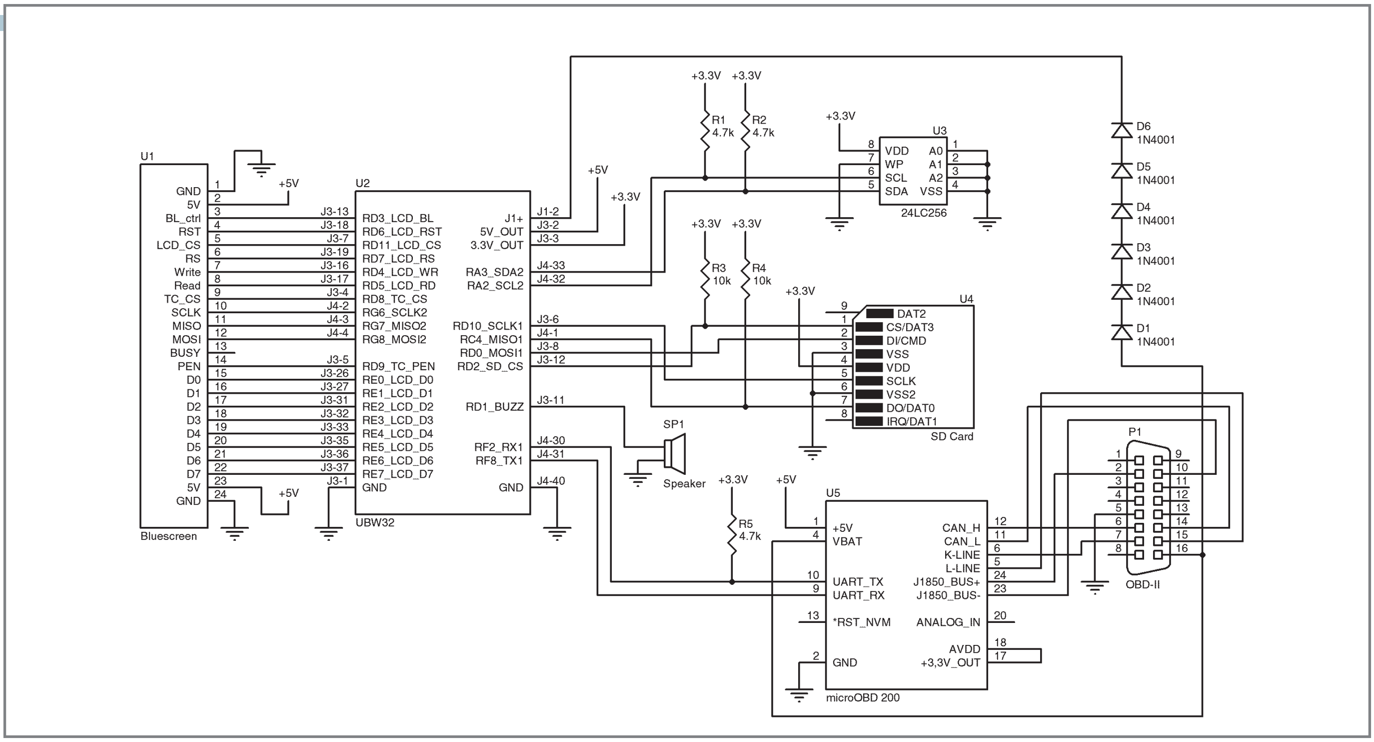


Figure 2—This schematic is for prototype testing. It has minimal filtering and protection. A later version will include a fuse and an LC filter.

as well as all the discrete components necessary for interfacing to the vehicle OBD-II diagnostic connector. Even though OBD-II is a standard, each manufacturer is free to use any electrical interface they desire, as long as the protocol follows the standard. This means you need different wiring to accommodate the various interfaces, such as VPW, KPW, and CAN. Along with different electrical wiring there is also different software involved. The microOBD takes care of all these headaches and makes every car, no matter what the manufacturer is, appear identical to the application program. The UBW32 connects to the microOBD with a TTL serial port. The application program talks to the vehicle by sending PID requests to the engine control unit (ECU). The ECU then responds with the desired data. For example, if you want to know the vehicle speed, send the string "`01 0D`" to the microOBD. The microOBD translates the request, sends it to the vehicle ECU, waits for the ECU response, translates it, and sends a response such as, "`41 0D 32`" to the application program, where "`32`" is the scaled vehicle speed.

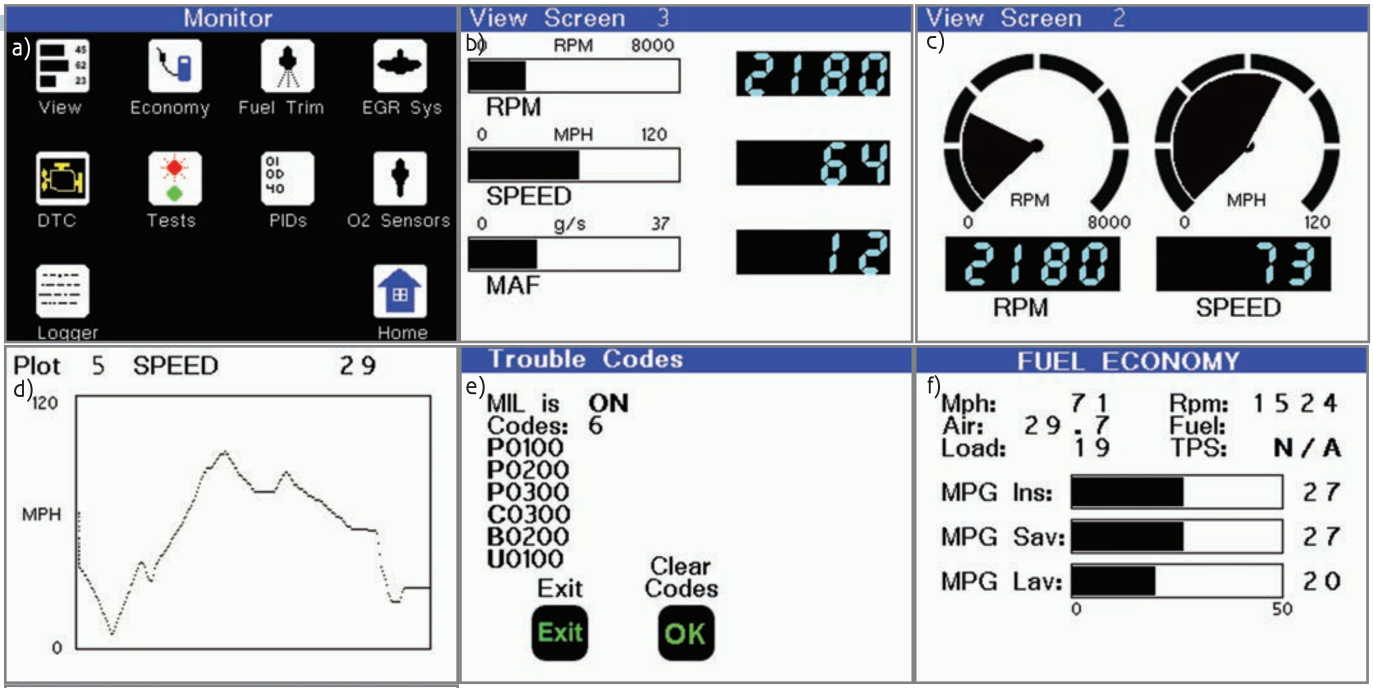
A SparkFun Electronics 256-Kb I²C EEPROM is used for nonvolatile storage of various configuration settings. The interface to the EEPROM is via an I²C port. An SD card socket is used for data logging storage. A handful of resistors, diodes, and a SparkFun buzzer round out the required components.

The 3.3- and 5-V power regulators on the UBW32 are used to power the entire project and the 5-V regulator is powered from the 12-V vehicle power through the OBD-II connector P1. Feeding a 5-V linear regulator with 12 V is a recipe for heat generation if there is much current involved. There are several diodes used to drop the 12-V vehicle power to around 8 V for feeding the 5-V regulator on the UBW32. The display has an LED backlight that can draw almost 100 mA at full brightness. At that kind of current, the 5-V regulator on the UBW32 got a little toasty when running at 12 V on the input, which is why the dropping diodes were added. A PWM output is used to drive the backlight so that the intensity can be varied, which helps lower the current requirements. A block diagram of the project is shown in Figure 1.

Warning: an automobile power system can be a very nasty environment. Alternator noise and high-voltage spikes are not uncommon. The schematic is for prototype testing and has minimal filtering and protection (see Figure 2). A fuse and some type of LC filter will be added in a later version.

PROTOTYPE ASSEMBLY

The prototype was constructed on a 4.5" × 6.25" piece of 0.1" perfboard with copper pads on one side. Standard wire wrap techniques were used to connect the modules. Photo 1a shows a picture of the top of the board and Photo 1b shows the bottom of the board. Header pins were soldered onto the display module and the UBW32, then the modules were soldered onto the perfboard. A standard 24-pin DIP wire-wrap socket was used to hold the microOBD module. Note that the SD socket was mounted on the bottom of the board so it could be soldered to the copper pads as a way to physically mount it. The pins on the SD socket are not exactly 0.1" on center, but



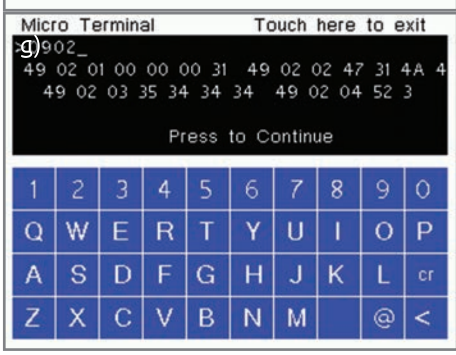


Photo 2—Screenshots of the menus, sample view screens, and some of the other tools

they were close enough to be soldered onto the pads. The OBD-II cable was hardwired to the board, but a cable with a DB9 connector is also available and could be used with a corresponding DB9 connector on the board, if desired.

SOFTWARE & OPERATION

The software is written completely in C and is compiled with Microchip's MPLAB compiler. It is a collection of pieces from different sources as well as my own work over several years of evolution. It's a constantly moving target. I am a self-taught programmer, so if you choose to use my code, please use it only as a guide, it is far from optimized.

The software can be broken down into a few main parts that correspond to the hardware modules: text and graphics functions, EEPROM read and write functions, OBD commands and responses, touchscreen scanning and coordinate calculation, SPI functions for the SD card interface, and a FAT file system.

The unit is set up with 16 user-configurable monitor screens and several preprogrammed screens, as well as some diagnostic tools and utilities. On each configurable monitor screen up to six variables can be shown, depending upon what type of screen is selected. You can choose the type of display parameters shown. There are five different types of monitor screens to choose from.

It was difficult to get good quality pictures of the various display screens for this article. They were either out of focus or they would have a glare spot. As a workaround, a utility function was written that captures the pixel data on the screen and writes it to a bitmap file on the SD card. Now the screen prints are crystal clear, which shows how versatile the hardware is. At this point, the screens are rather plain looking, which is probably due to the fact that I don't have a single artistic bone in my body. With a little effort and artistic talent they could be dressed up a bit. The border for the circular gages was actually drawn with Inkscape and converted to an .h file for displaying on the LCD, so there are possibilities for improvement.

One of the preprogrammed screens is used for monitoring fuel economy. It shows speed, RPM, load, mass airflow, and calculated fuel usage in miles per gallon (MPG). With the ever-rising price of gas, this screen comes in handy when trying to squeeze out a little more MPG.

Another screen gives you access to the malfunction indicator lamp (MIL) status. If the Check Engine light in your car comes on, you can use the scanner to retrieve the trouble codes and turn off the MIL, if desired.

There is a HyperTerminal-type utility that enables direct interaction with the microOBD module by issuing an AT/ST command and displaying the response. The scanner is also useful as a diagnostic tool. Watching the long- and short-term fuel trim values can help diagnose injector issues. Photo 2 shows screenshots of the menus, sample view screens, and some of the other tools.

WHAT'S NEXT?

Items on the to do list include adding a realtime clock/calendar backed up with a super capacitor, a mass storage device driver so log files can be easily transferred to a PC, a power manager to put the entire unit into low-power mode, and power supply protection and filtering.

In conclusion, using a modular approach, it is easy to construct a monitor for an OBD-II-compatible vehicle that enables you to scan many different engine functions.

Robin Brophy (myedash@yahoo.com) has a BSEE and an MSEE from North Dakota State University. He currently works in instrumentation and process control.

PROJECT FILES

To download the code, go to ftp://ftp.circuitcellar.com/pub/Circuit_Cellar/2011/254.

RESOURCES

Electronic Lives Manufacturing, FatFs Generic FAT File System Module, http://elm-chan.org/fsw/ff/00index_e.html.

Inkscape, Open Source Scalable Vector Graphics Editor, http://inkscape.org.

ScanTool.net, "User Guide ECUsim 2000 Multiprotocol Software Configurable OBD-II ECU Simulator," 2000, www.scantool.net/scantool/downloads/101/ecusim_2000-ug.pdf.

SOURCES

WinALDL ALDL reader
Jonas Bylund | http://winaldl.joby.se

ELM327 OBD-to-RS-232 Interpreter
Elm Electronics, Inc. | www.elmelectronics.com

Motorola MC68705 microcontroller
Freescale Semiconductor, Inc. | www.freescale.com

TFT Color LCD
Gravitech | www.gravitech.us

MAX7456 On-screen display (OSD) generator
Maxim Integrated Products | www.maxim-ic.com

PIC18F2620 Microcontroller
Microchip Technology, Inc. | www.microchip.com

BASIC Stamp microcontroller
Parallax, Inc. | www.paralax.com

microOBD 200 microcontroller and OBD cable
ScanTool.net | www.scantool.net

UBW32 USB 32-bit Whacker, PIC32MX795 development board, I²C EEPROM, and Buzzer
SparkFun Electronics | www.sparkfun.com

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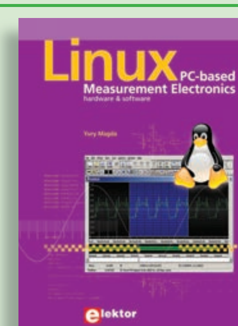

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Pico C Meter

Even upmarket digital multimeters boasting a built-in capacitance meter are useless if you want to check out tiny capacitances like 2.7 pF or 5.6 pF. Usually, you're tied to a lowest measuring range of 2000 pF, which is a good laugh to RF designers and radio amateurs. Although at 3.5 digits the DMM's resolution is 1 pF, any measurement below 200 pF or so will produce coarse if not ridiculous results. Pico C does a far better job. Beating many DMMs hands down, this little instrument easily measures capacitances down to fractions of a picofarad.

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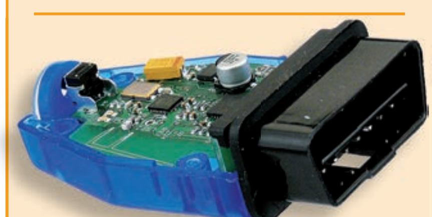
Art.# 100823-71 • $118.40



Wireless OBD-II

The cheapest way to diagnose faults on a modern car is to connect its OBD-II interface to a (notebook) PC running suitable diagnostics software. However, a wired connection is not always the most suitable, and self-contained OBD testers are a rather expensive and less flexible alternative to using a PC. An interesting option is a wireless OBD interface with a radio interface to a PC: this home-brew solution allows the choice of using either Bluetooth or ZigBee.

OBD-II-Zigbee or Bluetooth interface kit with all parts and enclosure

Art.# 100872-71 • $179.10 (Zigbee)

Art.# 100872-72 • $179.10 (Bluetooth)

Kits & Modules

by George Martin (USA)

Design Development (Part 3)

Software Design

The first and second parts of this series described a process for using C code and schematic-capture software to design a new product. Now it's time to delve deeper into software design—more specifically, the unified modeling language (UML).

A programmer, an engineer, and a philosopher were questioning who had the oldest profession. The engineer claimed his profession was the oldest because the creation of the heavens and earth was certainly an act of engineering. After a bit, the philosopher spoke up. His claim was that before the heavens and earth there was chaos, and out of that, order was created—an act of philosophy if ever there was one. After a bit more time had passed, the programmer spoke up and said, "Ah yes, but do you know who created that chaos?"

This is a great description of where we are in our product design—namely chaos. We've got some parts identified, some requirements seemingly pinned down, and a number of schematics left blank. You'll notice I have not drawn a schematic for the power supply, the USB/Ethernet interface, or the CPU. We don't know enough about those parts of the design to make any informed selections yet. So, I'm putting them off until later. This is a design technique I've seen described as the "*mañana* technique." *Mañana* is Spanish for "tomorrow." You have probably heard of "Top Down" or "Bottom Up?" Well, *mañana* puts off decision making. The theory is that as you work on the design more of it will be understood and better decisions can be made. I've been reluctant to talk about the *mañana* technique because it sounds like a poor (lazy) approach, but I find it works fairly well. Let's see how it unfolds with this project.

UNIFIED MODELING LANGUAGE

Just as we took the requirements and created schematics for each, let's take the schematics and the requirements and create a software design using Unified Modeling Language (UML). I've written about using UML before and there have been several articles in *Circuit Cellar* about it. Look them up. There's a lot of value in adopting these tools.

Starting with the temperature and humidity sensor, I created several routines (see Table 1). Well, here we are—more chaos. I tried to take the simple step of defining the temperature and humidity interface routines and I've got more questions than answers. Look at the return values for reading the temperature and humidity. They are integers. That implies temperature in degrees Fahrenheit but no decimal information. The temperature can go negative and UINT16 doesn't provide for that. Is temperature to the integer degree acceptable for this project? It's not in the requirements. Same for humidity, only integer values are reported. If this were a real project with a paying customer, I would start a

Routine	Passed parameter	Return parameter
`InitTHSensor`	None	INT16 Status
`ReadTemperature`	None	UINIT16 TempDegreesF
`ReadHumidity`	None	INT16 Humidity
`PowerUpSensor`	None	None
`PowerDownSensor`	None	None

Table 1—Unified Modeling Language routines

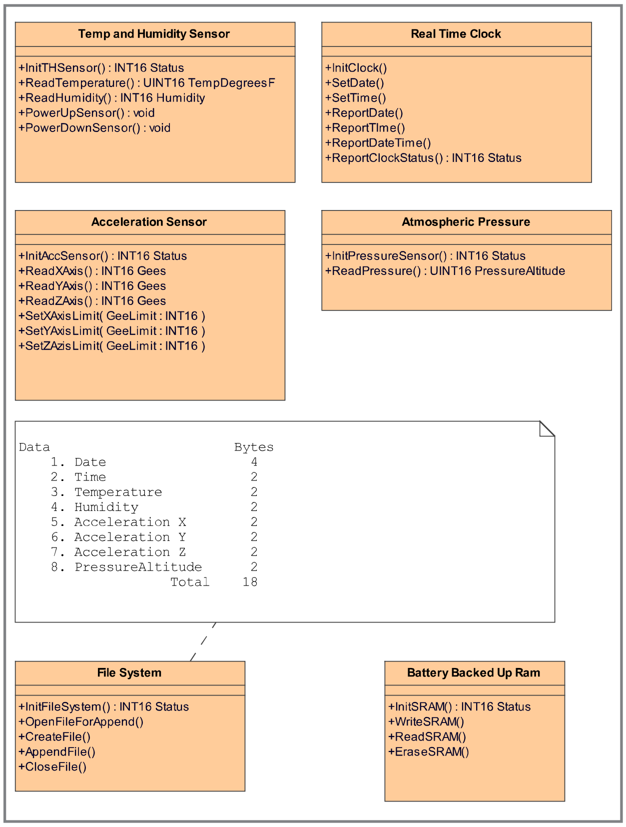


Figure 1—I used MagicDraw to generate planned operations (Unified Modeling Language diagrams)

list of questions and statements (or decisions) and put these two items on that list. If we need temperature to tenths of degrees Fahrenheit, now is the time to find this out. It affects the routine, the database, the reports generated, and probably a lot more.

If you've been reading my columns, you may have noticed that I have never talked about floating-point numbers in the C language. They do exist as data type in the language. But I find that a lot of code is required at run time to support them. If we needed to represent temperature in 0.1° it would probably require that we use floating-point routines. I'm going to try not to use them. Unless you really require support for floating-point variables, don't use them and you won't pull in the libraries and link to them.

We could represent temperature as TempInTenthsOfDegF and Humidity as HumidityInTenthsOfPercent and then save that in our databases in a binary format. The least-significant bit (LSB) would represent one-tenth of a degree. The reporting program would have to do the conversion. The reporting program is probably running on a PC or equivalent and has the horsepower to perform that conversion. So this might be an option if tenths of degrees are required.

Also, you'll notice the `PowerUpSensor()` and `PowerDownSensor()` routines. That's not in the requirements, but many of these sensors most likely have a Sleep or Low-power mode. If not, we could always add a high side switch on the power to the sensor. I bet the less power we use the longer we will be able to run, and that will make for a better product.

It's surprising how much we have uncovered without reading the component datasheets very closely. All we did was to create a place to write down the planned operations (UML diagrams) and we flushed out several issues (see Figure 1). We could have just started coding and run into these same questions at a later date, but that's no longer an efficient method of design.

REAL-TIME CLOCK

Let's take the same steps for the real-time clock (RTC) integrated circuit (IC). We need to initialize the clock chip and get a report of its status. This status should tell us if the clock has a valid date and if it has ever lost track of its time. Check the datasheet to see if those features are available. If not, I would move to another chip because our product needs this type of information. I have separate routines to set the date and time. It's a lot of typing for the user to enter the date and time in one command. With a lot of data to send, the process slows down and that means more delays. By separating them, you have less data to enter and less delay, so the clock setting more accurately matches the source setting.

Date and time are supported in the C language. There's much to read online. (Refer to the Resource at the end of this column.) I suspect there will be some time and data structures that eventually get described in our design. I usually keep copies of the RTC data in RAM, a time structure the C wants to work with, and routines that convert from one to the other. I don't think we need to get to that in this first pass of the software. You'll see more of this later on.

SENSOR ROUTINES

Remember, we have a three-axis accelerometer as part of the design. That device can report the acceleration in each of the three axes. But it can also generate an interrupt if the acceleration in an axis is above a set value. I created a routine to initialize the sensor and routines to read each axis. I also created routines to set the limits for each axis. I suspect we won't be able to know the orientation of the unit, so setting the limit will be one routine and the same limit will be set for each axis. The reading of all three axes could be done in one routine, but then I would pass a pointer to an array for the data and let the routine fill in the values. It may come to this, but until then, let's use the three separate routines I've defined.

For this sensor I've added an Initialization routine and a routine that reads the pressure altitude. Hopefully, this would be in feet above sea level. I'm not sure if we can do this, but let's try this approach for now.

NOTES & LOOSE ENDS

I added a note to our UML diagram listing the data that is to be saved. It's a first pass at the data we are gathering for each reading. I'm envisioning that each time we take a reading we'll collect and save all the readings.

There are several items missing. We haven't selected a microcontroller, looked at the USB/Ethernet interface, or done the power supply, and we don't have a memory card. And, I talked about an On/Off switch earlier, which we also don't have. Well, this is a series of articles on design development, and this is how I do the designs. We don't know enough about these missing topics to make solid decisions yet. That will depend on how much data we have to move about the system. Let's look at that next, and I bet several of these issues will clear up: the *mañana* design technique at work.

DATA RECORDS

One key area is how much data we are going to save. Let's start with a simple first-pass approach. From the notes in the UML diagram, I created a list of parameters (see Figure 2). For a first attempt, I want to make up a record that is all ASCII characters and easy for the human to read (see Figure 3).

Each variable is readable engineering units. We can save this data record to memory or print it out in exactly the same format. I'm not sure what units acceleration will be recorded as and I think pressure altitude should be in feet. I put the spaces in between each variable for the article. You could probably delete them in the data as it's saved and probably not make it too confusing. I suspect some readers may be wondering, why not pack the data into binary, as shown in the UML? Well, we can always do that, but for now, I believe it's easier for us humans to work with the data in this form. So, the data we save, print, plot, and transfer is all readable ASCII. Yes, it consumes more space, but for now, it will make debugging a whole lot easier. We can always use different bit-packing formats to save space.

How often should we record the data? The easy answer is: whenever it changes. But again, for a start, let's say every 6 s. That's 10 per minute and about 31 bytes per record. So 310 bytes per minute (i.e., 310 × 60 = 18,600 bytes per hour or 18,600 × 24 = 446,400 bytes per day). That does not seem like a lot of data, but as we get further into this design, we might be surprised. For now, this data record seems to work.

The acceleration is two decimal digits. The transducer has 6 bits of data per axis and we can select the measurement range to be ± 2, 4, or 8 g. That 6-bit resolution is ± 32 counts decimal and that 32 counts can represent 2, 4, or 8 g. So, there is a decision to make about the data record. Should we save the transducer reading (counts) or the

Parameter		Bytes
1.	Date	4
2.	Time	2
3.	Temperature	2
4.	Humidity	2
5.	Acceleration X	2 or 1
6.	Acceleration Y	2 or 1
7.	Acceleration Z	2 or 1
8.	PressureAltitude	2
	Total	18 Bytes

Figure 2—A list of parameters created from the notes in the Unified Modeling Language diagram

engineering units (g). Let's postpone that decision. For now, we'll just save the room for either. That probably means adding a + or – sign to the data array for each of the three axes.

The datasheet for Bosch Sensortec's BMP085 digital pressure sensor gives the range as "300 … 1,100 hPa (+9,000 m … –500 m above sea level)." If we're using English units, this would be about 30,000′ to –1,640′ above sea level. I'm suggesting we just record to the nearest 100′. So, 30,000′ becomes 300′ and –1,640′ becomes –16′. So, the three-digit format looks like it will work.

DATA TRANSFER & RECORDING

If we had to transfer one week of data, we would have 3,124,800 (i.e., 446,400 × 7) bytes of data to transfer. USB 2.0 can operate at 20 MBps, so USB is a candidate for communications. Ethernet (100BaseT) is 100 Mbps or 12 MBps maximum. Let's assume 50% is available for our transfer. So, that gives us a transfer rate of 6 MBps and our week's worth of data will transfer in less than 1 s. I'm not exactly sure what the speed requirements for transferring the data are. But less than 1 s for a week's worth of data should be acceptable. Remember, that's a new reading every 6 s. And, our data is printable ASCII characters; no data compression.

Another big part of the performance of the system will be reading and writing data to and from the memory card. Let's start by looking at the SD type of memory cards. I suspect these will meet our requirements and we will not need to go to the CompactFlash (CF) with its parallel data bus. Our week of data is 3.3 MB. A quick search online shows that a 2-GB card costs $8 and a 4-GB card costs $15. The smaller cards are starting to rise in price, so we should probably start with the 4-GB

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Specification
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NXP UDA1330 Stereo Audio DAC
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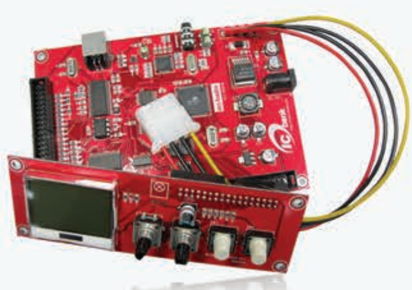

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Item	Specification
MCU	Atmel ATmega128L
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IDE Interface	Standard IDE type HDD(2.5", 3.5")
Power	12V, 1.5A
LCD	128 x 64 Graphic LCD
Etc	Firmware download/update with AVR ISP connector

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```
2011 JAN 01 11 59 59 -99 99 00 00 00 000 <CR> <LF>
 |    |   |  |  |  |  |   |  |  |  |   |   |    |
 |    |   |  |  |  |  |   |  |  |  |   |   |   <LF> Line Feed
 |    |   |  |  |  |  |   |  |  |  |   |  <CR> Carriage Return
 |    |   |  |  |  |  |   |  |  |  |  000 - Pressure Alt (000 to 999)
 |    |   |  |  |  |  |   |  00 00 00 - Acceleration in X Y Z
 |    |   |  |  |  |  |  99 - Relative Humidity %(00 to 99)
 |    |   |  |  |  | -99 - Temperature F (-99 to 200)
 |    |   |  |  | 59-Second (00-59)
 |    |   |  | 59-Minute (00-59)
 |    |   | 11-Hour (01-12)
 |    | 01-Date (01-31)
 | JAN-Month (JAN FEB MAR APR MAY JUN JUL AUG SEP OCT NOV DEC)
2011-Year
```

Figure 3—An easily readable record made up of ASCII characters

cards. Cards in capacity of up to 4 GB use the FAT16 format. You should read up on this to get an understanding of the issues. FAT12 was the first released followed by FAT16. As drives grow larger we need different formats to access all the drive space. Then we also have long file names to consider. We could spend several articles just talking about drives and formats. For now let's just stick with FAT16. We won't create or delete files using long file names but someone could have put a long file name of the memory card so we need to deal with them. The same issues are probably true for directories.

Let's also keep our file system simple by naming a file for the date. So, 3MAR2011.txt is the file we will write all the data into for the date of March 3, 2011. While the date is 3 March 2011, we check to see if the file exists. If it does, we simply append the data; if it does not, we create the file and the first data record. So, for a week of running, we'll have seven files (one for each day) that hold all the data. Each file will be 446,400 bytes in size or perhaps smaller. Changing cards on the fly can then be done anytime throughout the day.

Now we need to check to see if the file exists and, if so, append our new data to the end of the file. This operation will be time consuming (time consuming in our embedded time frame). I don't want to gather data and then write it to SD memory. Performing these tasks in sequence will take too much time. Remember, gathering data may include powering up the transducers and waiting for them to stabilize. So, I think we should have some memory that can hold the data even if the CPU goes to sleep. Then, gathering the data will be running in parallel with writing to SD memory. This approach just uncovered several additional requirements. We need two serial interface ports (I^2C or SPI), one port for the transducers and one port for the SD memory. We need memory that will hold data while the CPU goes to sleep. We need a file system with FAT16 support and possibly FAT32 support for the future. We need to run the two operations in parallel so an RTOS is probably in our future. Interesting how we flushed out these requirements by just designing with paper and pencil.

CONCLUDING THIS PHASE

I added the file operations to the UML diagram in the File System block. I also added what I thought would be routines we would need to interface to "battery backed-up RAM" and put that on the UML diagram. The UML diagram is starting to take shape. It's still very rough, but a design is starting to appear.

Next time, we'll look at processors and interfacing to the transducers in more detail. Until then, read up about support for time in the C language.

George Martin (gmm50@att.net) began his career in the aerospace industry in 1969. After five years at a real job, he set out on his own and co-founded a design and manufacturing firm (www.embedded-designer.com). His designs typically include servo-motion control, graphical input and output, data acquisition, and remote control systems. George is a charter member of the Ciarcia Design Works Team. He is currently working on a mobile communications system that announces highway information. He is also a nationally ranked revolver shooter.

RESOURCE

The University of Utah, School of Computing, "The GNU C Library Reference Manual," The GNU C Library, www.cs.utah.edu/dept/old/texinfo/glibc-manual-0.02/library_19.html, 1993.

SOURCES

BMP085 Digital pressure sensor
Bosch Sensortec | www.bosch-sensortec.com

MagicDraw Software modeling tool
No Magic, Inc. | www.nomagic.com

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Embedded Unveiled

by Richard Wotiz (USA)

"Embedded Unveiled" is a new bimonthly column devoted to embedded technology in real-world applications.

Linear Positioners

This column presents the inner workings of linear positioners—specifically media drives. Here you learn how exploring the linear electromagnetic actuator of a removable media drive breathed new life into an aging CD player.

Welcome to Embedded Unveiled! I'm excited about sharing with you an exploration into the inner workings of real systems. I'll be tearing apart perfectly good products to see, in great detail, how they work. This won't be anything like the static teardowns you may have seen elsewhere. I'll be showing the product in operation and discussing the theory behind some interesting aspect of its design. I'll be on the lookout for particularly clever or unusual technologies.

Much of what I learned about electronics when I was younger came from opening up off-the-shelf products and trying to figure out how they worked. It really helped me when it came time for a formal engineering education, since I found it much easier to understand the theory when I had already had some exposure to the practice. I hope this column can continue that journey for me and will inspire you to learn something new from an in-depth look at what others have designed.

This first column provides a detailed look at linear positioners. I'll be exploring the innards of removable media drives and how their magnetic or optical heads follow the data tracks on the media. Then I'll focus on a linear electromagnetic actuator that's used in a CD player. I'll cover the player's entire tracking system and show it in operation during various normal and adverse conditions. If you're wondering how I happened to choose this topic, read on!

I recently noticed that my home CD player wasn't working too well. It would occasionally skip when playing the higher numbered tracks on a disc. I shouldn't have been too surprised, given its age, since I had bought it more than 25 years ago in the days when CDs were still competing with vinyl records. But, the player seemed like it was built sturdily enough to last a long time, so I figured it was time to take a look inside and see if I could fix the problem. I took off the cover and realized it

Media	Tracks per inch	RPM (1× speed)	Run time (1× speed)	Tracking method
Phonograph record	150–300	33 1/3	15–30 minutes	Groove
Floppy disk	48–135	300 or 360		Stepper motor
CD	15,875	480–210	74 minutes	Linear actuator and voice coil positioner
DVD [1]	34,325	1,530–630	60 minutes	
Blu-ray disk [2]	79,375	1,957–810	90 minutes	

Table 1—Track geometries and positioning methods for some common removable media. The RPM ranges are for constant linear velocity (CLV) operation, where the drive starts at the maximum speed and gradually slows down as it progresses. Although the RPM and run times are for 1× speed, most optical disk drives run considerably faster.

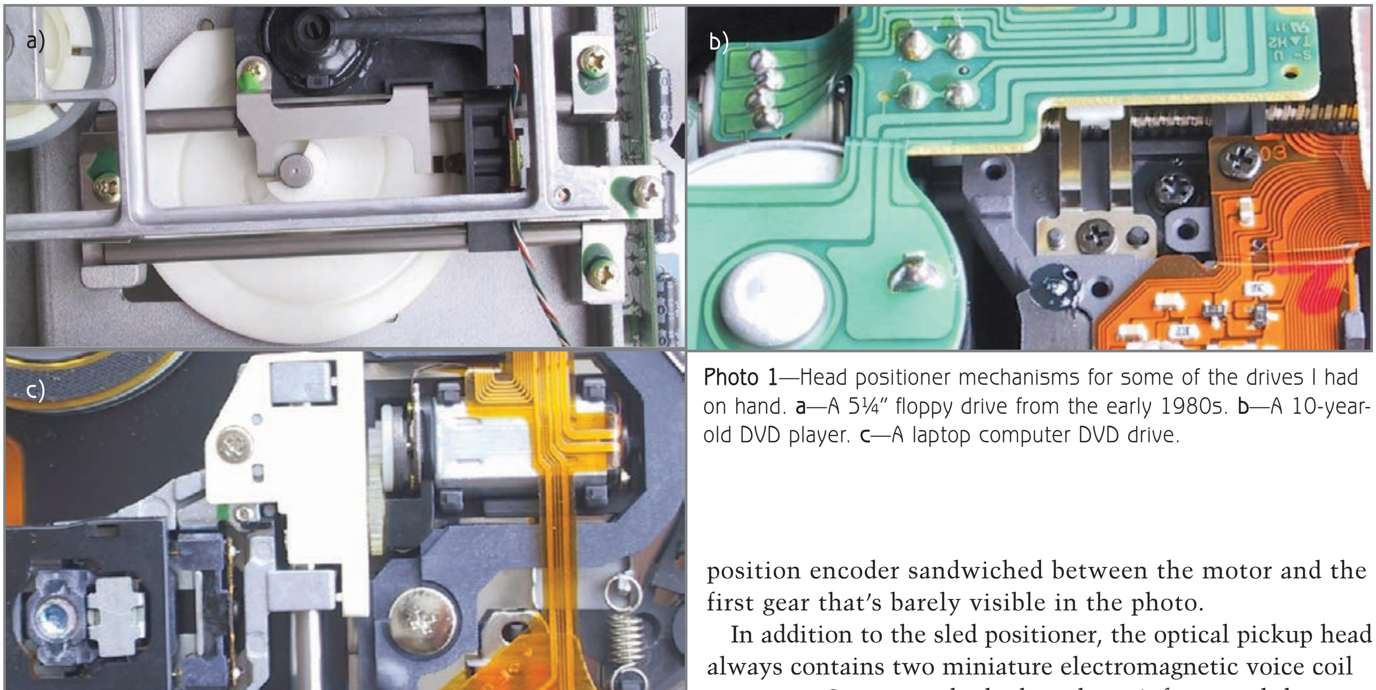

Photo 1—Head positioner mechanisms for some of the drives I had on hand. **a**—A 5¼″ floppy drive from the early 1980s. **b**—A 10-year-old DVD player. **c**—A laptop computer DVD drive.

had been a while since I had worked on anything so mechanically complex. As I tried to understand where the problem was, I became intrigued with how the tracking mechanism worked.

A HISTORY OF MEDIA

I was curious about how linear head-tracking systems for removable media have evolved over the years as data density has increased. Table 1 shows parameters for some of the more common media, and Photo 1 shows the head positioning mechanisms for several different drives. Floppy drives use an open-loop stepper motor to navigate between tracks. Their low track density doesn't require much precision, and a simple leadscrew or cam does the job. Photo 1a shows the mechanism of a 5 ¼″ drive from the early 1980s. If you look closely, you can see a spiral track on the large white gear that converts the stepper's rotary motion to a linear position to move the head.

The much higher track density of optical disks required a change from open-loop steppers to a closed-loop servo tracking system. A typical CD or DVD drive has two different head positioning systems. The first is a coarse positioner that moves a carriage or sled containing the entire optical pickup assembly. This is similar to the mechanism in a floppy drive. It isn't accurate enough to zero in on a particular track, but it can move from one end of the disk to the other. Different drives implement this positioner in different ways. It can be a linear voice coil actuator, or it can be a DC or stepper motor. Photo 1b shows a stepper motor with a worm drive. The tiny motor is almost hidden under four solder blobs at the top left above the much larger spindle motor. You can see the angled threads of the shiny worm gear just below the upper printed circuit board (PCB). The mechanism in Photo 1c uses an ordinary DC motor at the upper right driving several gears. There's a position encoder sandwiched between the motor and the first gear that's barely visible in the photo.

In addition to the sled positioner, the optical pickup head always contains two miniature electromagnetic voice coil actuators. One controls the laser beam's focus and the other is the fine-tracking positioner. Both have springs so they will return to center when current is removed. The tracking positioner can move the optical pickup to the exact location of a particular track. It's much faster than the sled mechanism and can easily follow any slight variation in the track position as the disk rotates. But, its range is limited, so the sled first has to get the pickup head reasonably close to the desired track.

MAGNETIC FORCE

My CD player's drive mechanism uses a linear voice coil actuator as a sled positioner (see Photo 2). Figure 1 shows its operation in more detail. The actuator consists of a ferrous metal frame holding a pair of permanent magnets. The magnets create a magnetic field, known as a B field, with flux lines that exit at the North Pole,

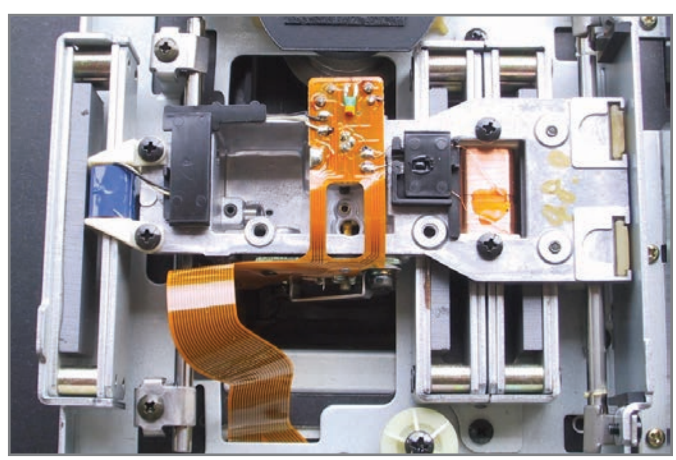

Photo 2—The CD player's sled is the horizontal structure that slides vertically along a track. The linear actuator's drive coil and its two bar magnets are toward the right. The small blue coil and single magnet at the left are the sled speed sensor. The optical head isn't visible here, it's mounted on the opposite side of the sled.

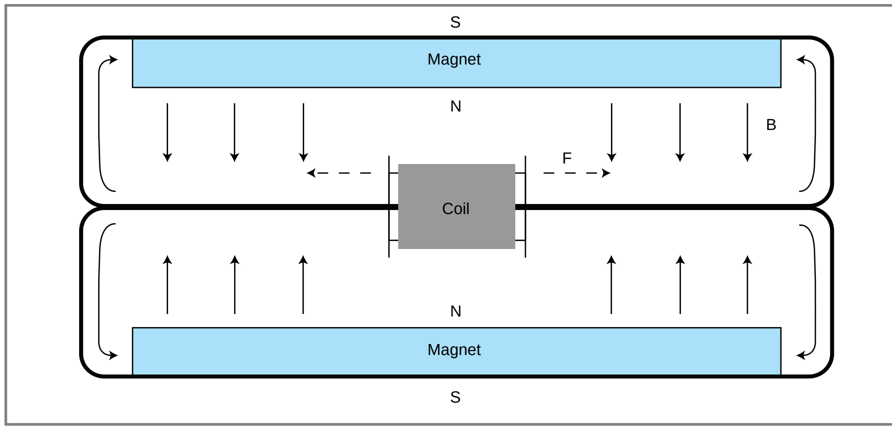

Figure 1—Top view of the linear actuator. The solid arrows are magnetic flux lines, which point from north to south. The dashed arrows show the force exerted on the coil. The direction of the force is determined by the polarity of the coil current.

cross the air gap to the center part of the frame, and then follow the frame back around to the South Pole. A drive coil that's attached to the sled straddles the center rail. Current flowing through the coil will cause it to move in either direction, depending on the applied polarity. The movement results from the Lorentz force, which is a force on a current-carrying conductor when it's in a magnetic field. The force is always perpendicular to both the current flow and the B field. When a voltage is applied to the coil, the current flowing in the segments of wire that are perpendicular to the B field will result in a force on the coil parallel to the center rail, as shown by the F arrows. This moves the sled in the desired direction.

Dutch physicist Hendrik Lorentz first quantified the Lorentz force in 1892. He determined the formula that represents the force on a charged particle due to electric and magnetic fields: $F = q [E + (v \times B)]$, where F is the force vector in newtons, q is the particle's electric charge in coulombs, E is the electric field vector in volts per meter, v is the particle velocity in meters per second, B is the magnetic field vector in teslas, and x is the vector cross product. When applying the formula to a coil in a magnetic field, it can be simplified to: $F = NL (I \times B)$, where N is the number of turns of wire, L is the wire length in meters, and I is the current vector in amps. Linear actuators that depend on this force have been around for more than a century.[3]

TRACKING CONTROL

Figure 2 shows a block diagram of the tracking servo loop. A pair of

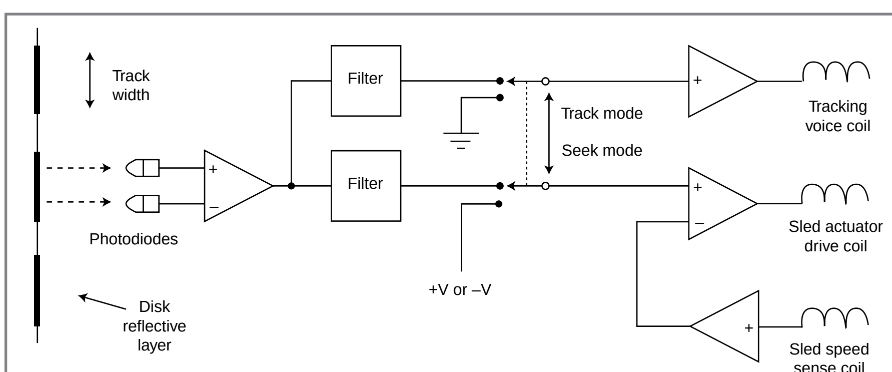

Figure 2—The CD player's tracking control loop. Any difference in illumination of the two photodiodes causes the drive coils to push the optical pickup to minimize the difference. The concept is remarkably simple for such a precision system. Some more recent players use a DSP in the sled control loop, but the operating theory is still the same.

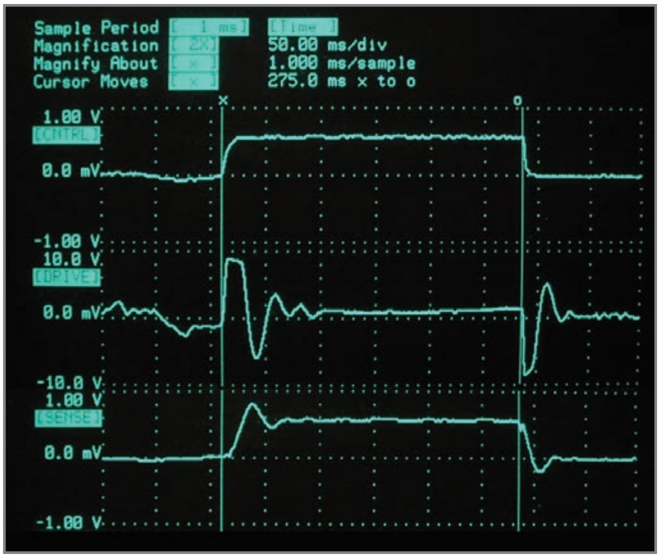

Photo 3—These are the waveforms from the sled actuator's drive op-amp during a seek operation. The top and bottom traces are the positive and negative inputs and the center trace is the coil current. The current probe was set to 5 mA/V, so the coil current waveform is actually 50 mA full scale. This is a composite photo, since my HP1631D can only capture two traces at once.

photodiodes on either side of the optical pickup head's centerline will produce a voltage with a polarity and amplitude that depend on how far off the head is from the center of the track. This error signal is amplified, filtered, and fed back to both positioners to keep the head centered on the track. The sled mechanism has enough friction that any small corrections are handled exclusively by the tracking voice coil. But, when the tracking coil approaches the end of its travel, the error signal will become large enough that the sled will start to move as well.

Once the sled is in motion, it takes very little force to keep it going. If a constant force were applied, it would continue to accelerate and overshoot the desired position. The CD player prevents this by measuring the sled speed and applying it to a feedback loop that controls the current fed to the drive coil. The physical configuration of the speed sensor is much like the actuator, but it uses only a single magnet (see Photo 2). As the sensor coil moves through a magnetic field, Faraday's law of induction tells us that a voltage will be induced in the coil that's proportional to its velocity.

The track is one continuous spiral, so the positioners are in constant motion to follow the track. However, when the player needs to skip to a different section of the disk, the tracking system switches to an open-loop seek mode. The tracking voice coil is disabled, and the player's control logic feeds a constant voltage to the sled positioner's driver. Photo 3 shows the signals involved in this process. I had the CD player jump from the beginning of the disk to a point about halfway through. The top trace shows the control signal from the player's microcontroller. You can see the sled coil's drive signal (middle trace) immediately jump up to its positive clipping limit. As the sled starts to accelerate, the bottom trace shows the sense coil voltage start to rise. When it reaches the same level as the control voltage, the drive signal starts to drop. The system eventually stabilizes with the drive signal barely above zero. That's just enough to compensate for friction while keeping the sled at a constant speed. When it's time for the sled to stop, the same process happens in reverse. But you can see that it takes less time to slow down than to speed up, since friction helps with the process.

Now that I had measured what the sled was doing, I wanted to see if I could apply Lorentz's formula to the system. I don't have a magnetic field strength meter, but I was able to measure all of the other parameters to varying degrees of accuracy. Then I made a rough estimate of the field strength. First, I needed to find a value for F. My rusty memory from high school physics recalled a couple of useful formulas: F = ma and v = at. I measured the sled position before and after the seek shown in Photo 3 and

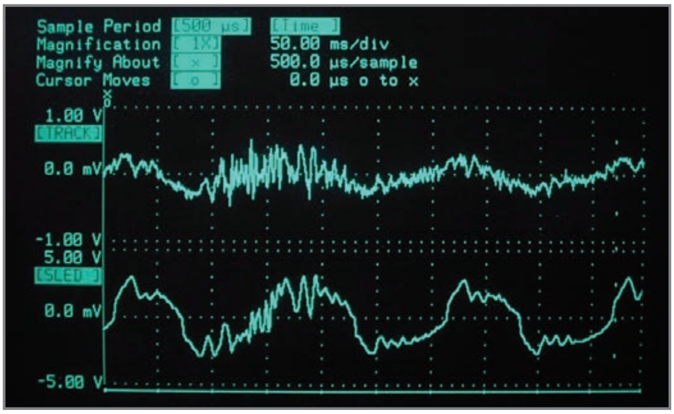

Photo 4—The tracking voice coil and sled positioner drive signals while the player is being bumped. It was a small nudge, so the tracking system was able to compensate for the movement.

Photo 5—Here the player is being bumped rather firmly. The tracking system can't keep up and loses synchronization for about 200 ms before it recovers.

found it had moved 17 mm in 265 ms, or a velocity of 0.064 m/s. Then, I zoomed in on the leading edge of the waveforms, and measured 13.5 ms for the acceleration from standstill to the steady-state velocity (prior to the initial overshoot). That's an acceleration of 4.74 m/s^2. To find the sled's mass, I precariously perched the CD player on its side and managed to press a small postal scale against the sled. The measurement varied from about 70 to 130 g, so I chose 100 g as an approximate value. That means we have a force of 0.474 N. Phew!

Measuring the coil was much easier. The wire diameter (with insulation) is just under 8 mils, which would be AWG 33. At 206 Ω per 1,000′, the coil, which I measured at 200 Ω, would have 971′ of wire. I measured the coil's circumference to be approximately 2.4″, which would yield 4,855 turns. The I vector from the Lorentz formula corresponds only to the vertical wire segments, which I estimated at 20 mm on each side, or 40 mm/turn. That gives a value of NL of 194 m. A close look at Photo 3 shows that the drive current during initial acceleration is about 40 mA. Now we have all the numbers we need! Plugging them all into the original formula and solving for B = F/(NLI) yields a value of 0.061 tesla, or 610 gauss. That's a little smaller than I thought I'd see, but still within an order of magnitude of what I'd expect for a ferrite bar magnet.

After all that work, I was curious to see how the player would handle a slight bump. Photo 4 shows the tracking and sled coil drive signals when I hit the table that the player was sitting on. The cyclical pattern follows the rotation of the disk, a result of the disk's center hole being slightly off center. The sudden increase in noise after one rotation is where the positioners are compensating for the bump. In this case, the system maintained tracking without any difficulty. Photo 5 shows a more severe impact that causes the player to momentarily lose tracking. You can see the sled bouncing around until it settles back down after about 1½ rotations. The player's audio dropped out for a short time during this process. Most newer players, especially portable ones, will buffer enough data to cover several rotations. They will also spin the disk at a faster speed, so it's possible to catch up after a short tracking loss without any dropouts.

PUTTING IT BACK TOGETHER

I'm happy to report that I was able to fix the CD player's skipping problem. As I moved the sled along its track by hand, it was obvious that one end was noticeably more stiff than the other. I carefully cleaned off all of the long-since-expired lubricant from the entire track and replaced it with a small amount of white lithium grease. That seemed to make more of a difference that I had planned. Now, even some CDs that wouldn't play at all—ones that I had assumed must have used some format incompatible with a 25-year-old player—play just fine.

I hope you've enjoyed this journey into the inner workings of a real embedded system as much as I have. I'll be taking a look at both conventional and exotic products in future columns. I'd also welcome suggestions for unusual products or technologies I may not be aware of. There will be a lot more to explore in upcoming months. I hope you can come along for the ride!

Richard Wotiz has been taking products apart ever since he was old enough to pick up a soldering iron. He's been helping others put them together since 1991, when he started his design consulting business. Richard specializes in hardware and software for consumer products and children's toys. He can be reached at rw601@spiraltap.com.

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[1] MPEG TV, "DVD Technical Notes," www.mpeg.org/MPEG/DVD, 1996.

[2] H. Bennett, "The Authoritative Blu-ray Disc (BD) FAQ," www.hughsnews.ca/faqs/authoritative-blu-ray-disc-bd-faq.

[3] E. Chaplin and H. Halloran, "Electromechanical Movement", U.S. Patent No. 575,699, 1897.

by Jeff Bachiochi (USA)

Lined Up with Nowhere to Go

Using a Linear Sensor Array

Want to know how a linear sensor array accumulates and transfers data? Here's an example of how the sensor can be used to read a barcode.

Each day brings new technology to center stage. Today's phones have higher resolution cameras than the digital camera we bought just a few years ago. Some phones even have two cameras. Dick Tracy's wrist radio is a clunky behemoth compared to smartphones. And now, your computer can be your phone. Or is that your phone can be your computer?

It wasn't that long ago that fax machines (remember those?) enabled us to send documents over the phone. That was a boon for businesses that, up to that point, had relied on snail mail to deliver documents. It was a marvel to behold. You just placed your document into the autofeeder and an optoelement broke the document into hundreds of lines of data, with each line consisting of hundreds of black or white pixels. The pixel data was streamed through a phone line to a receiving machine that would then replicate the original by printing the received sequence of pixel data for each line just as it was scanned.

Stepper motors pull the paper through the machine. The step size creates the vertical resolution, while an optosensor supplies the horizontal resolution. The optosensor's output might come from a single sensor that is stepped horizontally across the page or from a fixed linear sensor made up of hundreds of individual sensors. Using a linear array enables the line to be read in one fell swoop. Today's scanners/copiers still use the same approach (although they're able to detect color instead of just grayscale).

Texas Advanced Optoelectronic Solutions (TAOS) has some inexpensive linear sensors that can be purchased through some of the standard online and catalog sales channels. I chose the TAOS TSL1406R linear sensor array to experiment with for this month's project (see Photo 1).

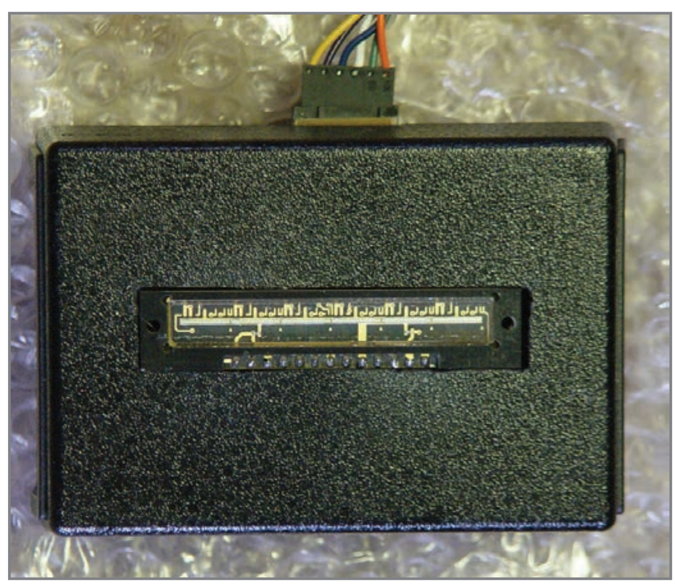


Photo 1—The TAOS TSL1406R is a 768 × 1 linear sensor array that has a built-in hold function that enables the last pixel samples to be temporarily stored while a new exposure is being taken. The stored analog samples are clocked out during the new exposure. The analog samples are therefore representative of the previous exposure period.

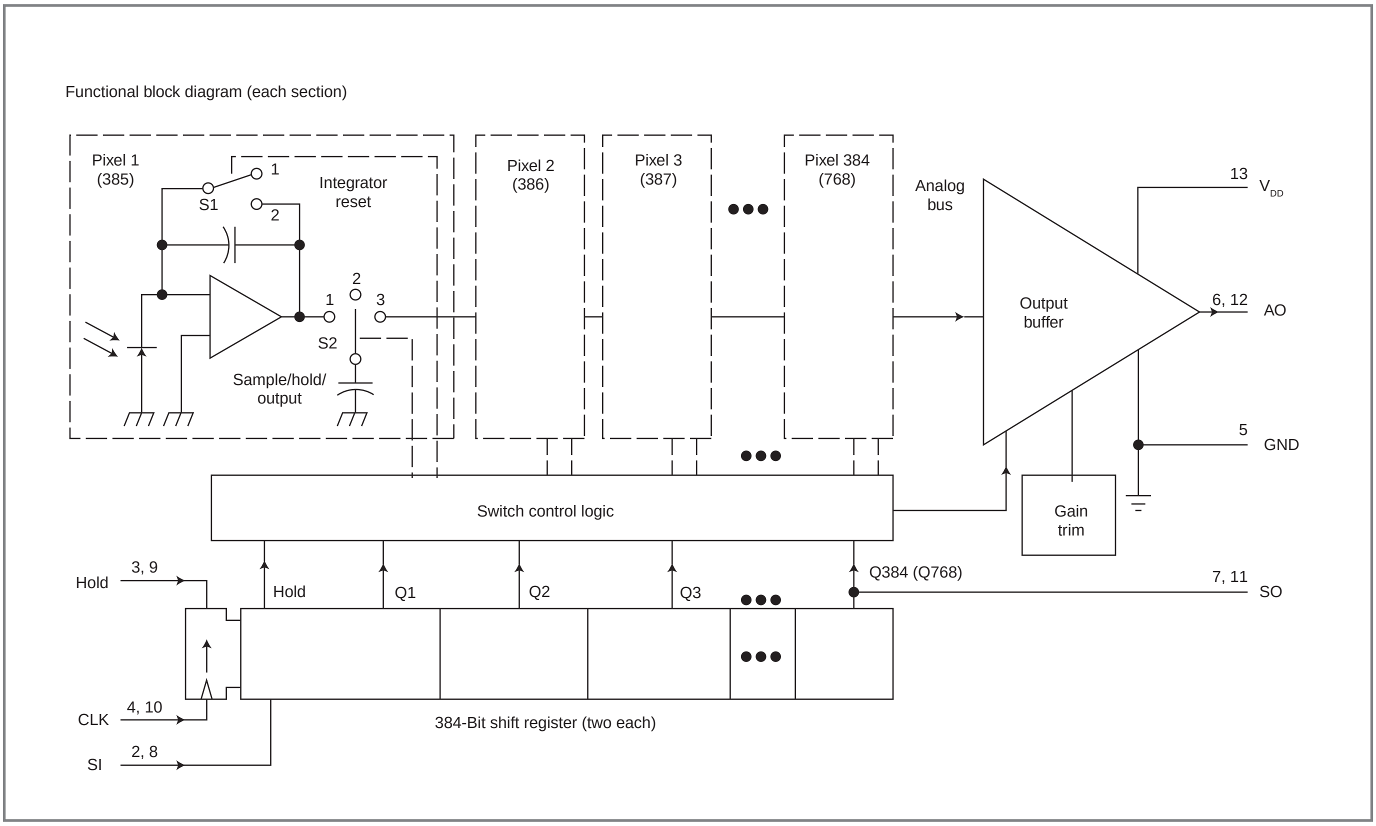


Figure 1—The TAOS TSL1406R linear sensor array consists of twin 364-pixel element linear arrays with individual sample-and-hold circuitry. A user-provided clock controls the exposure time while providing sequential control of the output of the previous exposure's analog samples.

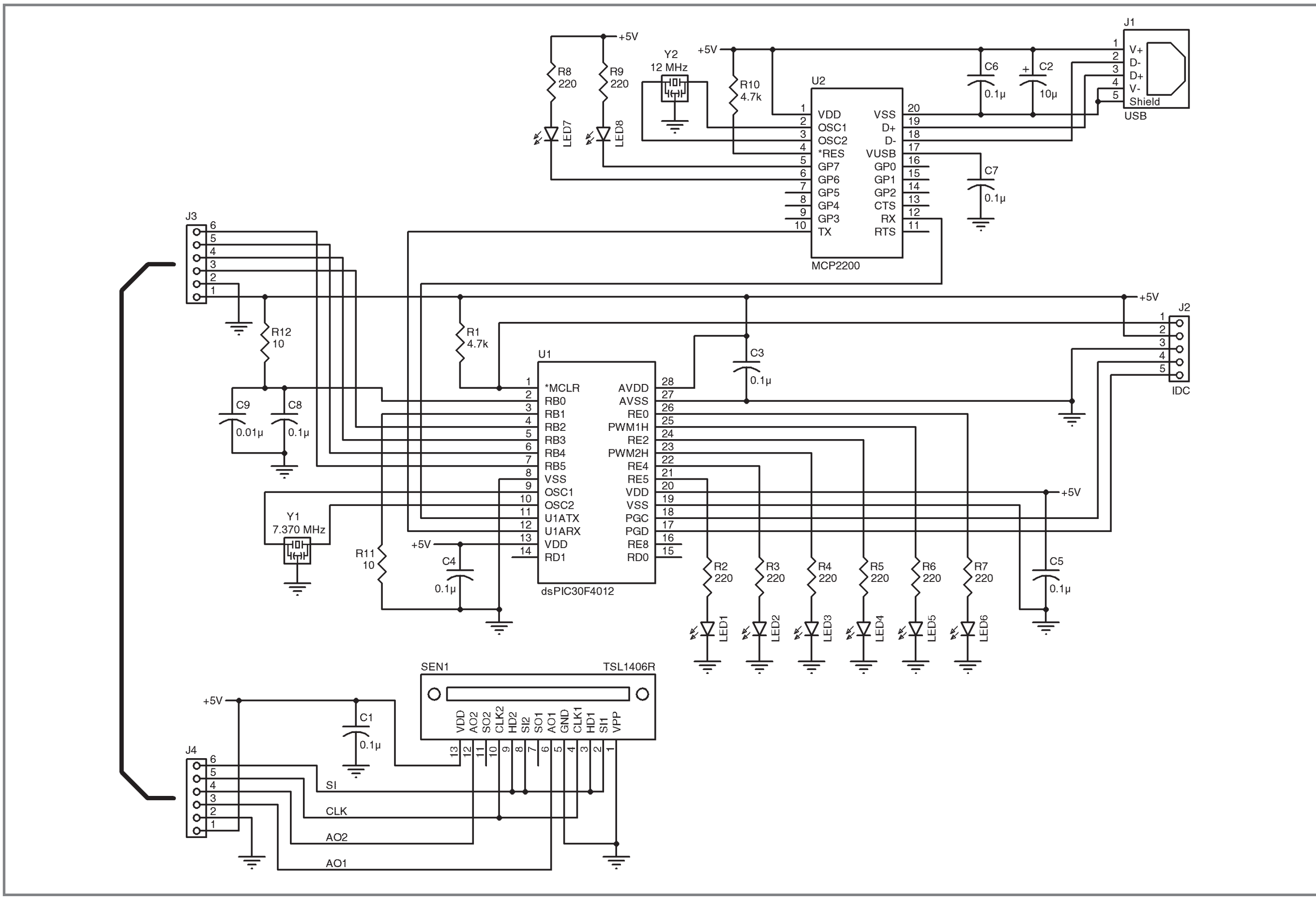


Figure 2—The Microchip Technology dsPIC30F4012 digital signal controller provides a maximum clock rate of 8 MHz to the TAOS TSL1406R linear sensor array for a minimum exposure time of 65.75 µs. The USB interface provides user I/O using an AT command set. LEDs are available for debugging purposes and provide some key timing information.

It has a 400 DPI resolution with a sensing area about as wide as a business card and costs about $25.

ELEMENTAL

Twin 384-pixel sensor arrays are set up end-to-end creating a single 768 × 1 linear element. Each 0.0025″ (63.5 µm) spaced pixel has its own integrator and sample-and-hold circuitry. This enables all pixels to begin and end a data integration phase in concert. The resulting integration values are stored for future conversion. These values are available for external sampling on a sequential basis. Two analog outputs are available, one from each array of 384 pixels. These can be sampled as separate arrays, in parallel, or cascaded using a single analog output.

As you can see in Figure 1, each array of 384 pixel elements has its own control I/O. The I/O connection determines whether each array works in parallel (simultaneously) or serially (independently). Three inputs, CLK, SI, and HOLD, control the entire integration process. In this project, the TSL1406R is used in Parallel mode with all SI and HOLD inputs connected together. It is driven from a single CTRL output and both arrays are clocking in step from a single CLK output (see Figure 2).

For this discussion, let's assume the array has already accumulated data and we want to read this data from the linear sensor. A logic-high pulse by CTRL for one clock period will start an output cycle. TSL1406R control inputs are sampled on the rising edge of the CLK. When the HOLD input is evaluated as a logic high, all sampling capacitors (holding integrated pixel data) are disconnected from their respective integrating circuitry. The integrating circuitry is reset and prepared for a new integration period. The same logic high on the SI input applies the first pixel's sampling capacitor to the charge-coupled output amplifier and output pins AO0 and AO1. The pixel data is available to the external sampling circuitry during a single clock cycle. Additional clock cycles shift the initial SI pulse through a 384-bit shift register, each stage enabling pixel data for a single clock cycle in sequence. During the 384th clock time, the shifted SI pulse has now reached the SO output. Note: When used in the Serial mode, the SO from the first array can trigger the SI to the second array. In our case, since the arrays are being clocked in parallel, the SO is not used. Our external sampling circuitry must actually perform two samples, from AO0 and AO1, at the same time. More on this shortly.

As you've probably guessed, the CLK period controls the integration period or the amount of time a pixel's light level is collected. The maximum clock frequency of the sensor is 8 MHz, which sets the minimum integration time as:

$$T_{INTMIN} = \left(\frac{1}{CLK_{MAX}}\right) \times (N_{PIXELS} - 18) + 20\ \mu s$$
$$= 125\ ns \times 366 + 20\ \mu s$$
$$= 45.75\ \mu s + 20\ \mu s$$
$$= 65.75\ \mu s$$

In order to enable the sensor to operate in a number of different light level situations we will need to lengthen the integration time. I'll do this by slowing the clock down. The datasheet states that $T_{int(max)}$ should not be greater than 100 ms for accurate measurements. Working backwards with the preceding equation we see that the CLK must not be less than:

$$\left(\frac{1}{CLK_{MIN}}\right) \times (N_{PIXELS} - 18) + 20\ \mu s = T_{INTMAX}$$
$$\left(\frac{1}{CLK_{MIN}}\right) \times (N_{PIXELS} - 18) = T_{INTMAX} - 20\ \mu s$$
$$\frac{1}{CLK_{MIN}} = T_{INTMAX} - 20\ \mu s$$
$$CLK_{MIN} = \frac{(N_{PIXELS} - 18)}{(T_{INTMAX} - 20\ \mu s)}$$
$$= \frac{366}{(100\ ms - 20\ \mu s)}$$
$$= 3{,}661\ Hz$$

Providing a 3,661-Hz clock should not be an issue. An 8-MHz clock, on the other hand, is a challenge. In fact, each CLK cycle actually has four phases. The CLK line itself will toggle twice for each cycle. And, we will need to set/clear the CTRL input in between CLK toggles (at least for the initial SI/HOLD pulse).

dsPIC30F4012

I'm using a Microchip Technology dsPIC30F4012 digital signal controller (DSC) because it has a phase-locked loop (PLL) to boost oscillator frequencies by a factor of 16. The maximum input to this device is 8 MHz (using the 16X PLL). I've chosen to go slightly less, using a 7.3728-MHz input for its nice data rate division. So, the F_{OSC} is about 118 MHz and an instruction cycle is four clocks, a quick 34 ns. Timer2 can use the PR2 register as a period or count-up-to-and-then-rollover register. By not using a prescaler, I can get an event flag every instruction cycle (34 ns) if PR2 = 1 or any multiple of 34 ns up to 2.2 ms (i.e., 65,535 × 34 ns). This covers the bases between 8 MHz to 3.6 KHz as a clock source for the TSL1406R. Take a look at Figure 3 to see how the CNTL and CLK outputs were created. The first three instructions of this routine create the CTRL pulse that begins the sampling process. By strategically placing these among some inline setup code, the minimum timing requirements can be met. Timer2 can then be used to signal the appropriate times for toggling the CLK. Note there are two other things happening during this time. Before the CLK returns to the high state, there is a check to see what mode we're in. When `FLAG.SAMPLE=0`, the `Call SAMPLE` routine is ignored. Otherwise, this is the point where the analog outputs AO0 and AO1 from the linear sensor will be sampled by the internal A/D converter in the microcontroller. After the CLK returns to the high state, you check to see if you have looped once for each pixel: if not, then you jump back for more, or else you can finish by lowering the CLK one last time.

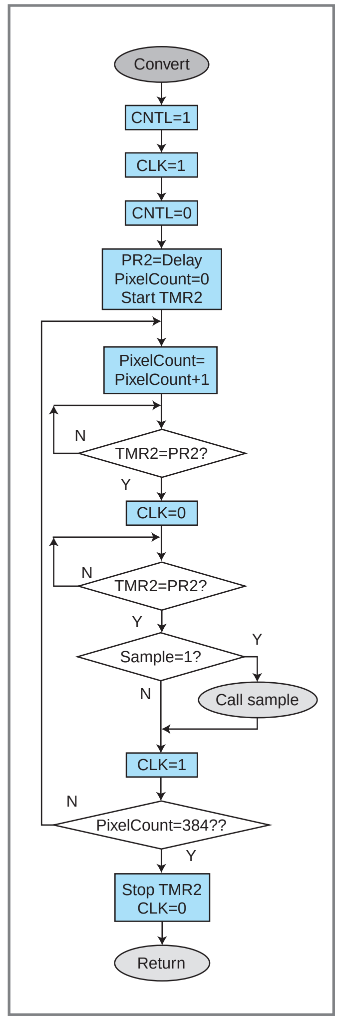


Figure 3—The convert process, a CTRL pulse during the first clock cycle followed by 384 additional clock cycles, is used to both expose the sensor and to transfer analog data. When `FLAG.SAMPLE=0`, there are no additional delays (other than the timing delay that sets the overall exposure time by slowing down the clock). When `FLAG.SAMPLE=1`, the data from the previous exposure is sampled by the microcontroller's ADC. Two analog outputs from the TAOS TSL1406R linear sensor are simultaneously sampled and then converted in succession.

The execution time for the loop while `FLAG.SAMPLE=0` is about 20 instructions or about 780 ns (i.e., 20 × 34 ns). So, any value of PR2 less than about 10 will not offer any faster loop time. While you can get an 8-MHz CLK by hard-coding a four-instruction loop, once code is added to vary the timing, the minimum time grows beyond 125 ns—in this case to 780 ns.

SIMULTANEOUS SAMPLING

The 10-bit A/D converter used in this microcontroller is fast, 1 Msps, but not fast enough to convert pixels on the fly at the maximum CLK frequency of the TSL1406R. It can, however, sample multiple (up to four) inputs simultaneously. You can take advantage of this since the linear sensor has two analog outputs. The actual conversion is handled sequentially after simultaneous sampling.

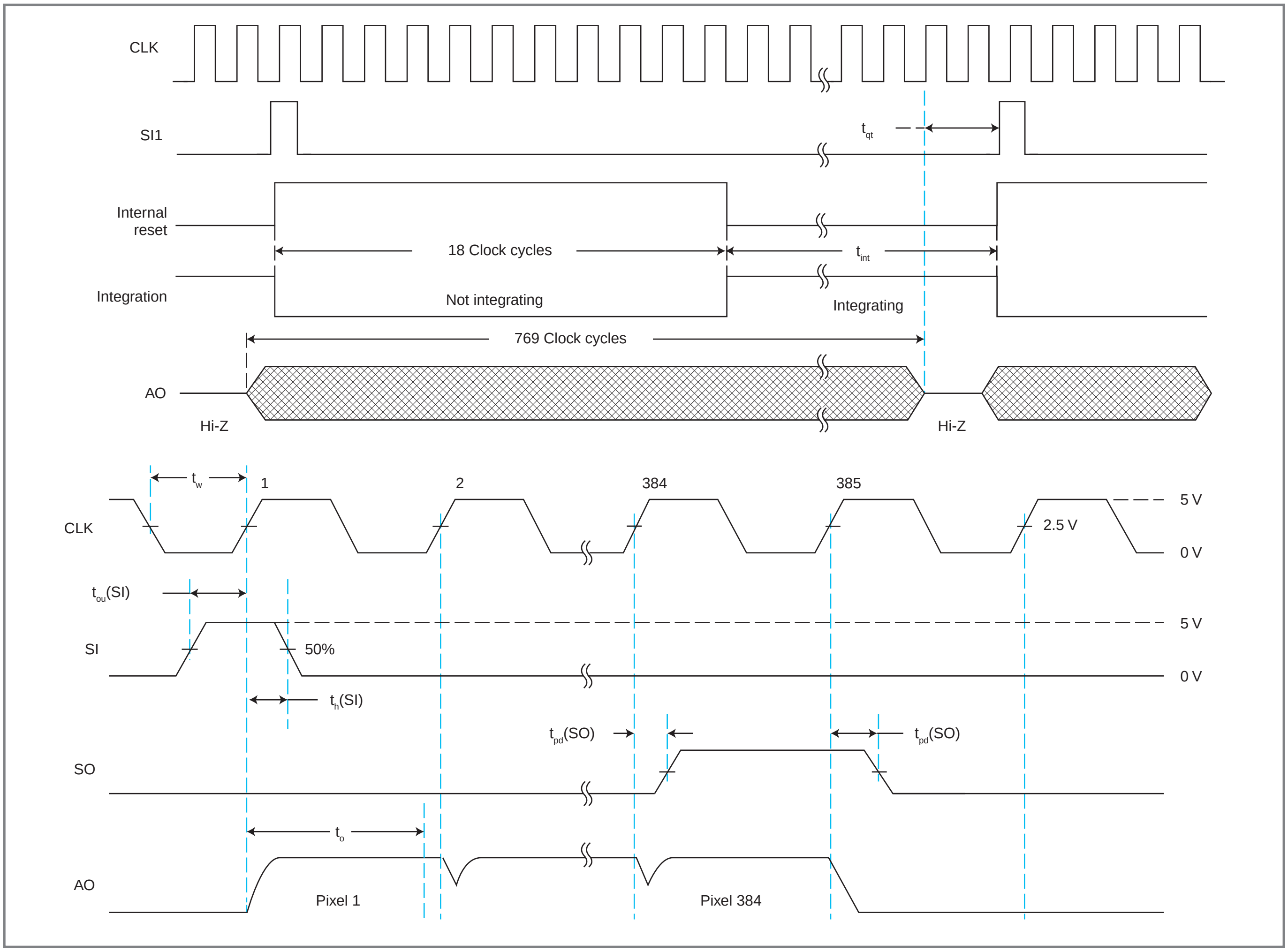


Figure 4—These timing diagrams of the TAOS TSL1406R linear sensor array show the relationship between the clock, the integration (exposure) time, and the analog output of pixel data.

Figure 4 shows the TSL1406R's timing and Figure 5 shows the microcontroller's A/D timing. Four independent sample-and-hold circuits can be instructed to take simultaneous samples of any analog input during the T_{SAMP} period. This can trigger automatic sequential connection to and conversion of each sample. Note: Once

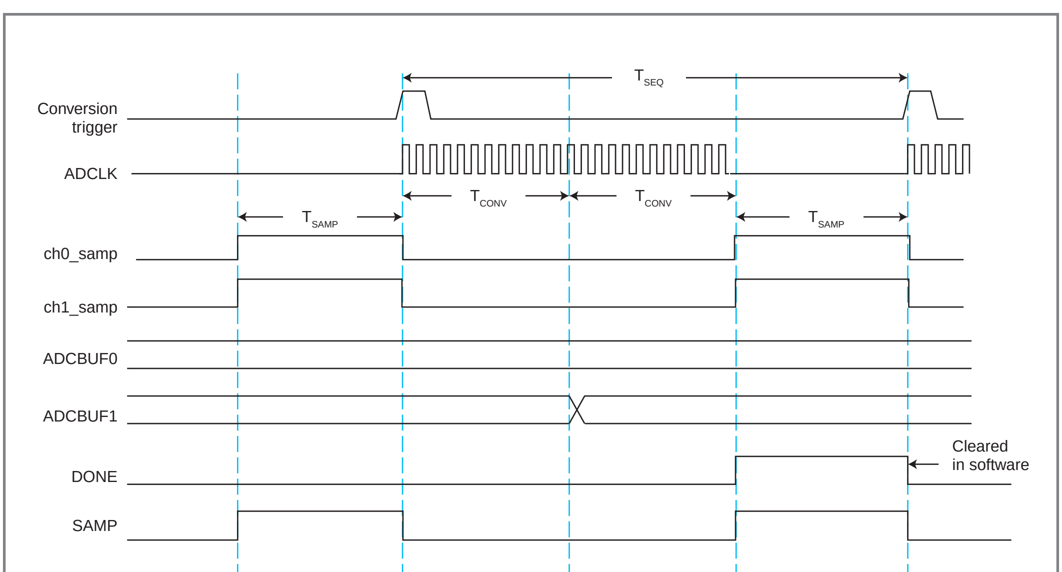


Figure 5—The timing of Microchip Technology's dsPIC30F4012 digital signal controller's high-speed ADC is set up to automatically convert channels after taking simultaneous samples of both of the TAOS TSL1406R linear sensor array's analog outputs AO0 and AO1.

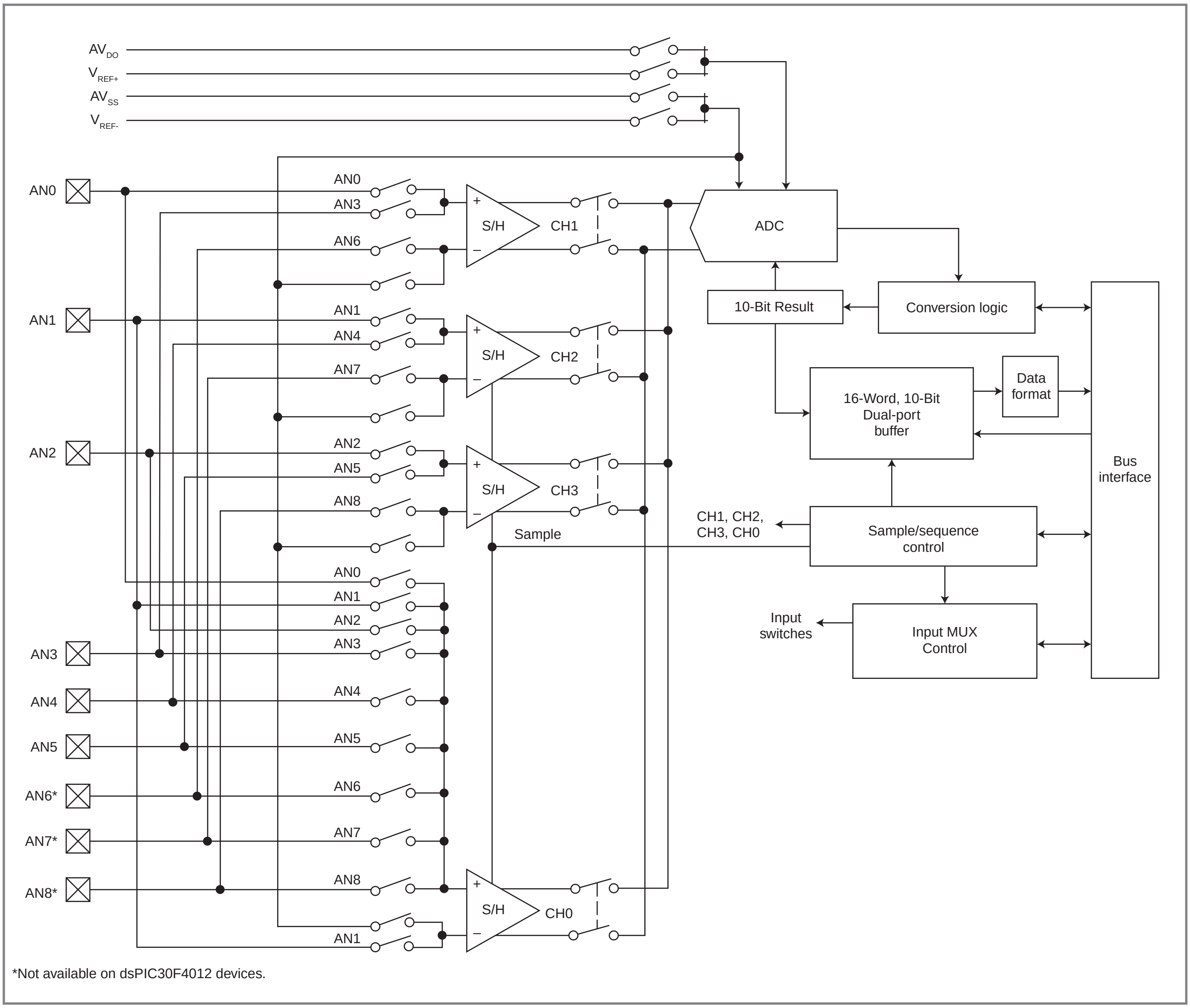


Figure 6—The high-speed ADC of the Microchip Technology dsPIC30F4012 digital signal controller can simultaneously sample up to four input channels. An alternate set of four channels can also be chosen. Automatic conversion of each channel can take place after the sampling period.

the T_{SAMP} period has concluded, the inputs are free to change (in this case, the next pixel output can be clocked).

The block diagram of the microcontroller's ADC shows a 16-word data buffer (see Figure 6). This can be initialized as either two eight-word buffers (ping-pong style) or a single buffer. Conversions are automatically dumped here and provide some flexibility as to when the data will be processed.

TSL1406R

If it wasn't already apparent, the linear sensor actually performs two operations during each cycle, integrating new pixel data and transferring old pixel data. In truth, the data on the analog outputs is from the previous cycle. After power-up, the initial data during the first output cycle is bogus. To simplify things, when I want to retrieve data for a particular exposure (CLK speed), I perform two output cycles. The first accumulates the data and the second transfers the data.

Refer back to Figure 1. Each of the 768 pixels in this linear array is a light-sensitive photodiode. The diode produces a voltage that is directly proportional to the amount of light that falls on the pixel. Each photodiode serves as an input to an op-amp circuit configured as an integrator. The integration circuit creates a rising voltage over time, which is proportional to the diode current. During the first 18 CLKs of an output cycle, a switch shorts out the feedback capacitor to discharge the integrator. During the remaining CLKs, all integrators continue to accumulate a charge. In fact, this continues until the next SI pulse begins a new output cycle. The accumulated charges are transferred to their respective sampling capacitors and wait their turn to be sampled via AO0 and AO1.

You can see that the total integration time is based on the CLK frequency plus any delay in starting another output cycle. As previously stated, this plan requires

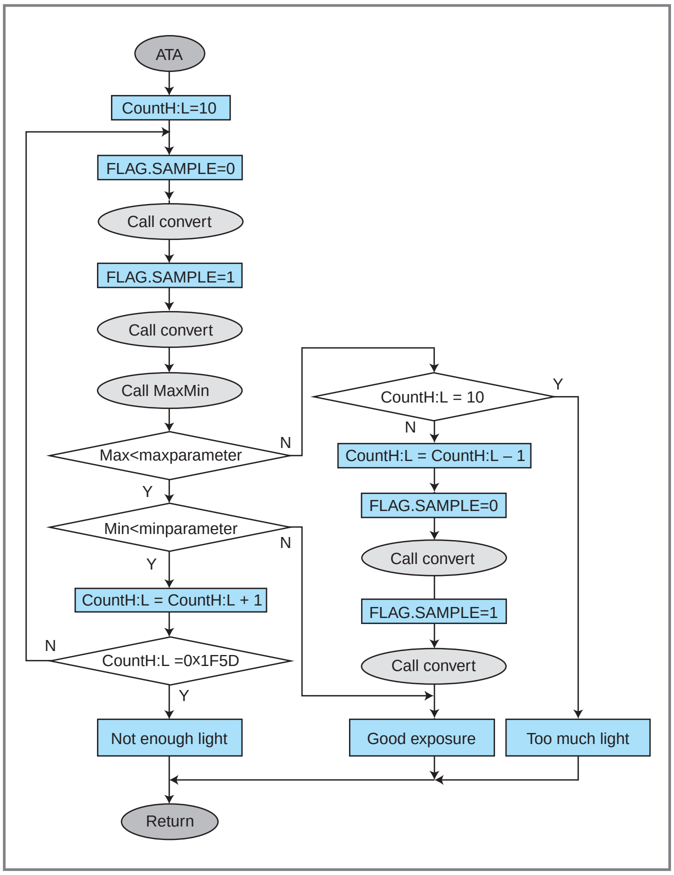


Figure 7—This flowchart shows how the ATA command performs multiple exposures (calling convert), beginning with the shortest and progressing in 32-ns increments. If all 768 sample values of an exposure fall within user-settable maximum and minimum limits, the routine exits with sampled data in the SampleBuffer.

two output cycles to actually retrieve data. The first output cycle uses stringent CLK timing to obtain a consistent integration rate. During the second cycle, A/D conversions are performed between each CLK. If the A/D sampling and conversion time is longer than the requested integration CLK rate, the total integration time for the second cycle may exceed the initial request. Therefore, using two output cycles always ensures accurate capture.

MICROCODE

I couldn't think of a good way to display the 768 values of data with a small number of LEDs or even an LCD. Because the data values are best viewed in relation to one another, I chose to dump the data out of the serial port. At a minimum, using a serial terminal would enable the user to scroll through the data and get a "feel" for what the linear sensor was seeing.

I created an `AT` command set because I was coming off the OBD-II project that used this format. I only needed a few commands: `ATS` (do two output cycles using the present integration count), `ATD` (dump the data out the serial port), `ATRC` (read integration count), and `ATWCxxxx` (write integration count).

The integration count is the hexadecimal value that gets loaded into the PR2 register of Timer2. If you recall, TMR2 increments every 34 ns. When the value in TMR2 reaches the value in PR2, an event is triggered. This enables us to create events (delays) from 34 ns to 2.2 ms. The toggling of the CLK output is based on these events. Assuming the consecutive output cycle takes place without any delay, the minimum integration time of 65.75 µs will depend on an 8-MHz CLK. This requires a TMR2 delay of 0x0002 (i.e., 2 × 34 ns = 68 ns or half of 0.125 MHz). Remember the minimum loop time is about 20 execution cycles, so any value of PR2 less than 10 won't have any effect. The maximum integration time is 100 ms, or a CLK of 3.6 KHz. This requires a TMR2 delay of 0x1F5D (i.e., 8029 × 34 ns = 273 µs or half of 1/3,661 Hz). So, the practical limits for PR2 would be 0x000A for a minimum integration time and 0x1F5D for a maximum integration time.

Figure 3 shows a routine `SAMPLE` that gets called when `FLAG.SAMPLE=1`. `SAMPLE` is set to 1 before the second output cycle. This enables the microcontroller's ADC to sample the pixel values on outputs AO0 and AO1 at the end of each CLK (after the data has had time to settle).

With flexibility comes complexity. The most difficult part of the ADC is determining which of the many modes of operation to use. For this project, there are four analog inputs (including +/–REF), no automatic scanning or alternate channels. Both sample and conversion timing are minimal. Two channels are being simultaneously sampled using external references. Conversions are automatic while sample start is triggered.

To begin a sample `ADCON1.SAMP` is set. When the sampling is completed, the peripheral clears `ADCON.SAMP`. Each channel is sequentially converted by the ADC with the conversions placed into `ADCBUF0` and `ADCBUF1`. The peripheral now sets `ADCON1.DONE` and is ready to perform the next sample and conversion. After each set of conversions, the most significant 8 bits of each value are moved into the 768-byte linear buffer. Since corresponding pixels from AO0 and AO1 are actually 386 pixels away from each other (linearly), the sampled and converted values of these need to be placed accordingly.

A data dump consists of sending a string for each pixel.

Photo 2—A close-up of the individual lenses in a strip of SELFOC lens array (SLA) held above the surface of a barcode photo. When affixed properly between the linear array and object to be read, it focuses the image of the object upon the array's linear elements.

The string takes the form of a three-digit pixel number and a blank followed by x periods, where x equals the converted value of that pixel. This enables the length of characters in the string to be a visual indication of light level.

RUN TIME

After some minor debugging, it was clear that the sensor's large dynamic range meant manually setting an integration count for each sample was a hassle. The integration time needs to be adjusted based on the converted data. That is, the integration time needs to be shortened so pixel data doesn't reach the saturation level (approximately 4.8 V). It also needs to be lengthened if no data reaches the white condition (approximately 2 V). Should the integration time need to be adjusted beyond its maximum or minimum times, the conversion is considered illegal.

To accomplish this automatically and maintain some flexibility, I added a few more commands: `ATA` (perform multiple two-output cycles to look for an exposure where all data falls within maximum and minimum parameters), `ATRX` (read maximum parameter), `ATWXxx` (write `xx` to maximum parameter), `ATRN` (read minimum parameter), and `ATWNxx` (write `xx` to minimum parameter).

The method used to automatically find an exposure is not sophisticated. Refer to Figure 7. Multiple pairs of output cycles are executed with the exposure time increased each time. After each set, all 768 sensor values are read to find the maximum (white) and minimum (black) values for the set. If the whitest value has exceeded the maximum parameter, you check to see that you are not at the minimum exposure (10). If you are already at the minimum exposure, then you are done and flagged as "too much light." If you're not at the minimum, then the exposure is reduced to the last value, reconverted, and flagged as a "good exposure."

If the whitest value remains less than the maximum parameter, the minimum value is checked against the minimum parameter. If the minimum value of the set remains less than the minimum parameter, the exposure is incremented. If the exposure has reached the maximum count (0x1F5D), then you're done and flagged as "not enough light (available)." If the minimum value (black level) is greater than the minimum parameter, you're done and flagged as a "good exposure."

BRINGING IT ALL INTO FOCUS

Each of the linear sensor elements will report on the amount of light that falls on it. A thin optical glass cover protects the elements (see Photo 1). In a transparent mode, any object placed between the light source and the sensor will block light from those elements directly behind it. However, the further away the object is from the sensor elements, the more light can bleed behind objects, partially illuminating the pixels behind it. Of course, the size of the object will also affect its shadow. The sensor also can be

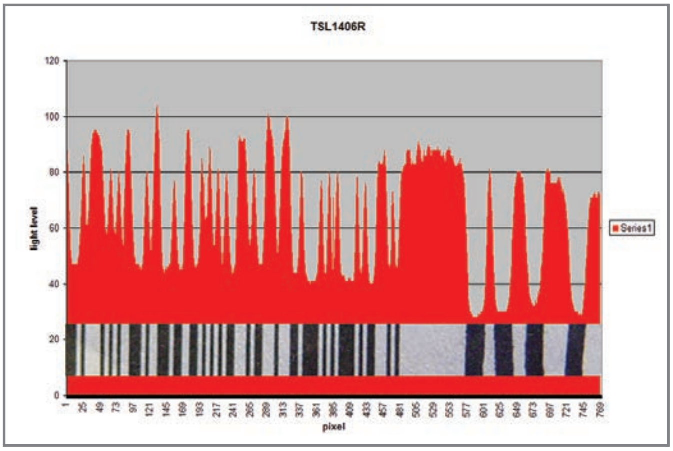


Figure 8—The sensor data using Excel. The insert is a slice of the actual barcode I read using the TAOS TSL1406R linear sensor array. I stretched the barcode to scale it to the graph's data.

used in a reflective mode. An example would be to sense the different reflective properties between white paper and black ink or toner.

In a fax/copier, a lens is used to focus the object onto the sensor. Special SELFOC lens arrays (SLAs) reduce the optical path length, giving a 1:1 focused image without the inversion produced by simple lenses. Photo 2 is a close-up image of a double-row SLA held slightly above a barcode. Note the holes at each end of the linear array in Photo 1. These mark the centerline of the sensor elements aiding in lens alignment.

I wasn't sure what to expect when I tried to read a barcode, so I placed a few wide magic marker stripes next to it. My terminal program captured the data output from the sensor and I imported it into Excel so it could be graphed. Figure 8 shows the pixel array's output graph. I took a picture of the barcode strip I used and superimposed it on the graph after stretching it to match the data. You can clearly see the relationship between the red graph area's illumination levels and the white background of the barcode.

While I succeeded in conquering the electrical operation of the TSL1406R in this project, I can't help but feel there is much more that could be done with this sensor. I look forward to receiving some input from you on how this sensor may be used to solve a real-world problem. I purposely did not touch on some issues, such as spectral response. These may be of importance if the application requires wavelength relationships.

One more note. The first time I powered up the circuit and ran some code, I quickly removed power as the microcontroller seemed to be running a lot warmer than most of my PIC projects. After searching for some miswiring that could create a short somewhere, I realized I was running the little PIC at close to its 120 MHz (30 MIPS) limit. Speed comes with a cost of approximately 170-mA operating current. While it looks like it's just sitting there, it's really screaming inside!

Author's note: Special thanks to Walter Boyles of GoFoton for this information on SLAs.

Jeff Bachiochi (pronounced BAH-key-AH-key) has been writing for Circuit Cellar *since 1988. His background includes product design and manufacturing. You can reach him at jeff.bachiochi@imaginethatnow.com or at www.imaginethatnow.com.*

PROJECT FILES

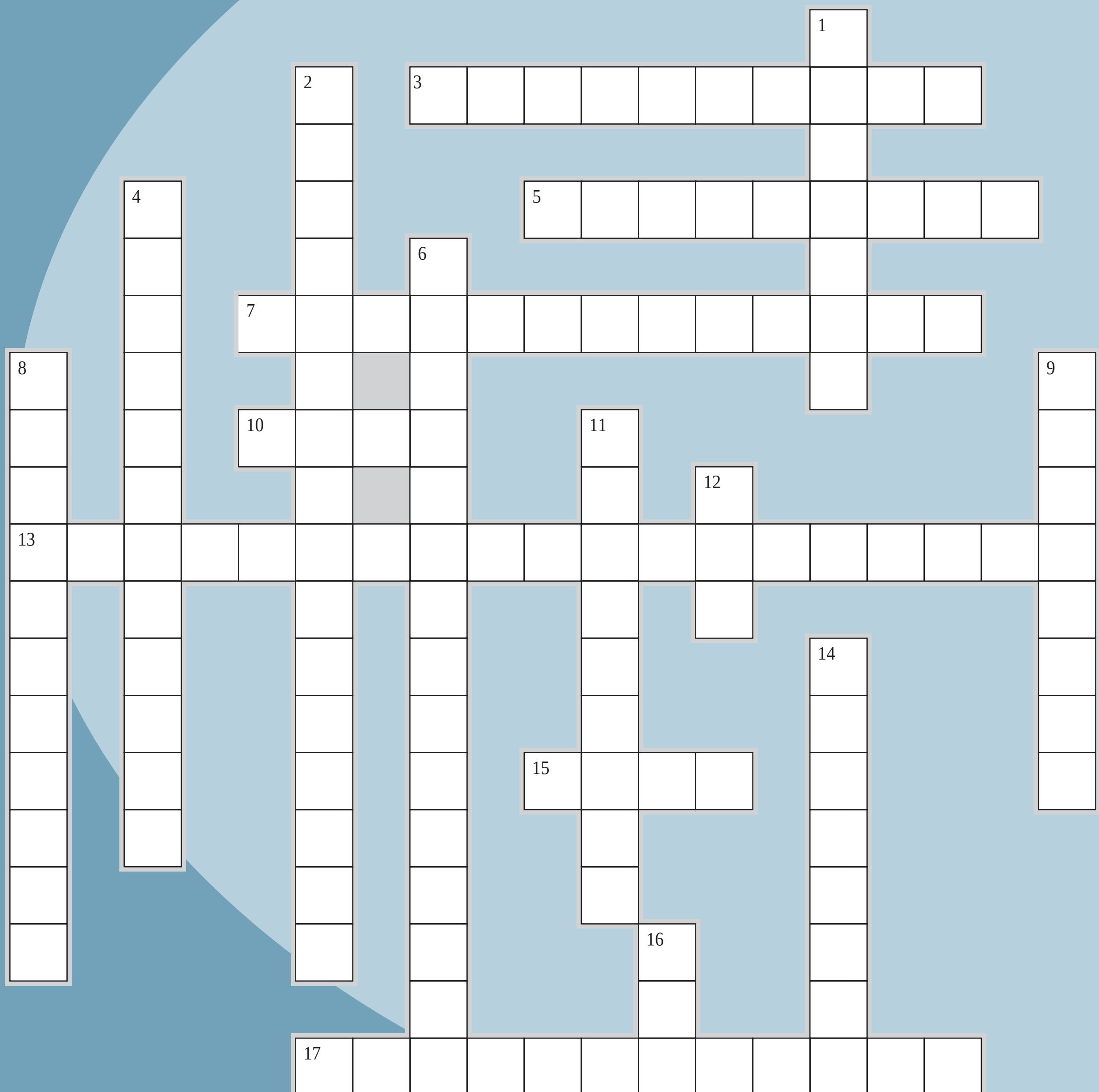
To download the code, go to ftp://ftp.circuitcellar.com/pub/Circuit_Cellar/2011/254.

SOURCES

dsPIC30F4012 Digital signal controller
Microchip Technology, Inc. | www.microchip.com

TSL1406R Linear sensor array
Texas Advanced Optoelectronic Solutions, Inc. (TAOS) | www.taosinc.com

CROSSWORD



Down

1. Automated, computerized
2. Helps you learn how to use a new microprocessor [two words]
4. Makes things seem less distorted
6. Topic featured in this month's *Circuit Cellar* [two words]
8. A message-based standard for cars [two words]
9. The physical side of your computer system
11. Once characterized by a rainbow-striped apple
12. Tagged image file format
14. Disconnected
16. A power supply that should not be disturbed

Across

3. Portable way to update your Facebook status
5. A wireless communication standard used to transfer data short distances
7. A system that represents real numbers in a variety of values [two words]
10. Field-programmable gate array
13. PCB [three words]
15. An SI unit
17. Reads positioning [two words]

The answers will be available in the next issue and at www.circuitcellar.com/crossword.

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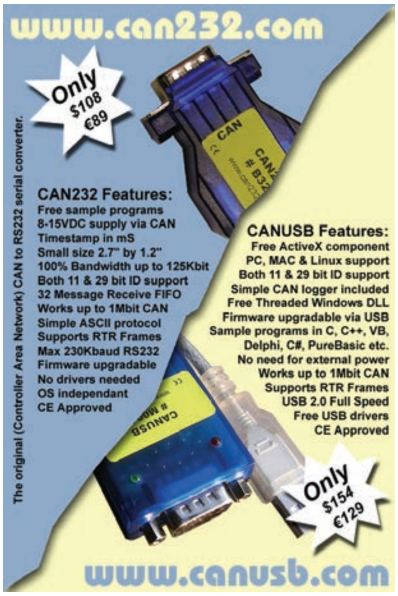
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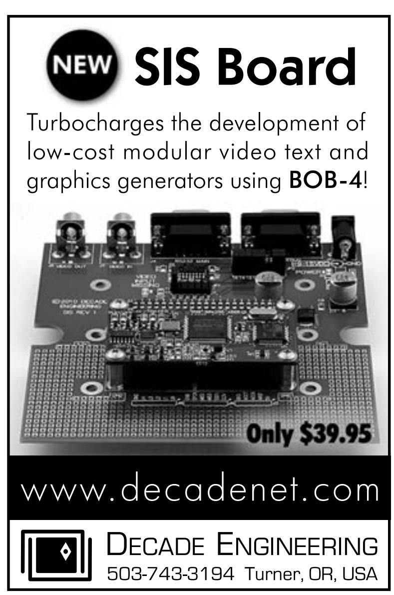
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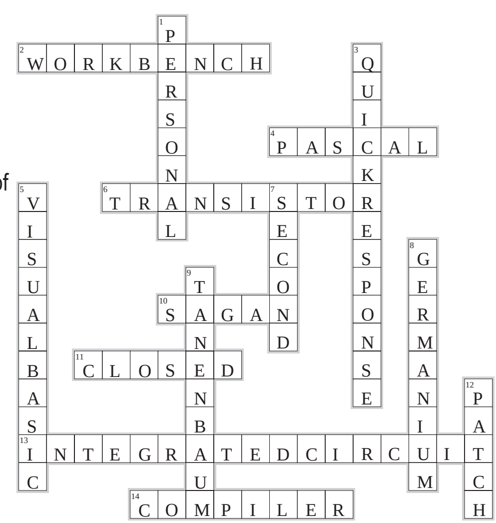
CROSSWORD ANSWERS from Issue 253

Across

2. WORKBENCH—The place to accomplish things
4. PASCAL—Pa
6. TRANSISTOR—Patented by Julius Edgar Lilienfeld in 1925
10. SAGAN—Slang for a massive quantity; think: Carl
11. CLOSED—Not public (e.g., software)
13. INTEGRATEDCIRCUIT—555 Timer [two words]
14. COMPILER—Needed for C, C++, Java, etc.

Down

1. PERSONAL—"P" in PIN
3. QUICKRESPONSE—Ge; semiconductor
5. VISUALBASIC—A derivative of BASIC [two words]
7. SECOND—Hertz = one cycle per?
8. GERMANIUM—Ge; semiconductor
9. TANENBAUM—Authored MINIX
12. PATCH—A quick fix



INDEX OF ADVERTISERS

PREVIEW of October Issue 255

Theme: Signal Processing

High-Accuracy Voltage Reference Using Pulse-Width Modulation (Part 1): Pulse-Width Modulation Theory

PID Temperature Control: The Ultimate MCU-Based Coffee Cup Warmer

Embedded System Communication: A Control Platform for Ethernet-Enabled Systems

---

EMBEDDED IN THIN SLICES **Getting Started with Embedded Linux (Part 1):** When to Choose Linux for Your Embedded Design

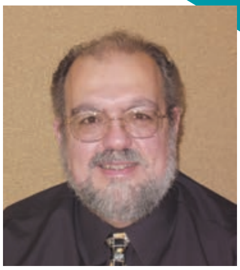
THE CONSUMMATE ENGINEER **Environmental Stress Screening**

ABOVE THE GROUND PLANE **Stepper Motor Drives:** Microstepping

THE DARKER SIDE **Line Coding Techniques**

FROM THE BENCH **Get Your Head Out of the Cloud:** Access an API for Communication with a Wi-Fi Thermostat

Priority Interrupt



by Steve Ciarcia, Founder and Editorial Director

Power Factor Correction

Last month's editorial about my problems getting a UPS that worked on my desktop hasn't been in the hands of readers for more than a couple days and I'm already getting e-mails from people experiencing similar problems. Obviously, I hit a resonant chord among the computer literate, but simply identifying the problem didn't do enough to explain power factor or the possible real cause for my failed UPSes. This month I'd like to fill in some of those blanks.

Basically, the term "power factor" is used in two completely different ways these days. The first way involves the traditional reactive load, where the voltage and current waveforms are both sine waves, but there's a phase shift between them. The numerical "power factor" in this case is the cosine of the phase angle. For example, a lightly-loaded motor appears very inductive; consequently, in industries that use lots of them, power factor correction capacitors are placed in parallel with the motors to compensate for that and avoid a surcharge from the power company.

For example, consider a 120-V, 1-A resistive load versus a 120-V, 1-A reactive load that has a power factor of 0.6. Both of them have a rating of 120 VA, but the first one also consumes 120 W of real power, while the reactive one only consumes 72 W (i.e., 120 × 0.6) of real power. Yet, in the second case, the power company's distribution network still has to carry the full 1 A of current, even though they only get to charge for the effective 0.6 A that delivers real power to the customer. That 1 A causes the same $I^2 \times R$ losses in the distribution network in either case.

That was sufficient back in the days when 99% of the load on the power grid was either resistive (heating, lighting) or motors. However, especially since the development of large amounts of consumer electronics, another source of excessive loss for the power grid has been identified: nonsinusoidal current waveforms, a.k.a. "harmonic distortion." The biggest culprit by far is the classic capacitor-input switching power supply, which draws all of its load current at the peak of the AC voltage waveform. This highly distorted waveform also causes excessive losses for the power company relative to the real power they can charge. This is because the RMS value of the distorted current waveform—which corresponds directly to the $I^2 \times R$ losses in the grid—is much higher than the RMS value of a sine wave current delivering the same amount of power.

This gives rise to the other usage of the term "power factor," which is completely different. There's no well-established definition for the numerical value of power factor in this case, but one that's often used is the ratio of the RMS value of the sine wave current to the RMS value of the actual current waveform. For example, this is how the Kill-A-Watt power meter measures power factor—it measures actual power, RMS volts, and RMS amps and then divides watts by (volts × amps) to come up with a power factor number. (I can only presume that it doesn't sample fast enough to properly characterize the amount of harmonic distortion in my power supply properly.)

However, that doesn't really give you the whole picture when it comes to UPS ratings. A typical medium-power capacitor-input power supply—regardless of whether it's a switcher or not—draws all of its current during maybe 10–15% of the AC cycle. The peak current would then be on the order of 10× what you would think based on the power rating alone. In other words, a 240-W load could be drawing current peaks on the order of 15 to 20 A! But the power meter will show this as a power factor of about 0.45 to 0.55. If you're still thinking in terms of reactive loads, you may decide that you need a UPS rated at 533 VA (i.e., 240 W/0.45).

But the UPS doesn't really care about RMS values—it has to deal with the instantaneous load current at all times. For example, if your UPS uses 24-V batteries, the 20-A peak on the output translates to a current of more than 100 A on the primary side. The UPS is monitoring the instantaneous current and the smaller ones shut down in self defense as soon as they see this current spike, which is well over their ratings. In other words, the VA rating of a UPS is simply there to give you an idea of its peak current capability, not its ability to handle reactive loads. A 480-VA UPS can deliver 4 A RMS at 120 V, or 5.66 A peak. A 1,500-VA UPS can deliver 17.7 A peak, and now we know why the large UPS was required to power the 240-W load.

In order to avoid having to buy way-oversized UPS units, you need to look for computers and monitors that include "active" power factor correction (PFC) in their power supplies. This is a circuit that redistributes the current draw back to a normal sine wave shape, basically by adding another switching circuit upstream of the main regulator. Apparently, my desktop doesn't have this feature, but you can learn more about PFC in Dave Tweed's EQ quiz (*Circuit Cellar* 214, 2008) www.circuitcellar.com/eq/214/214.html.

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