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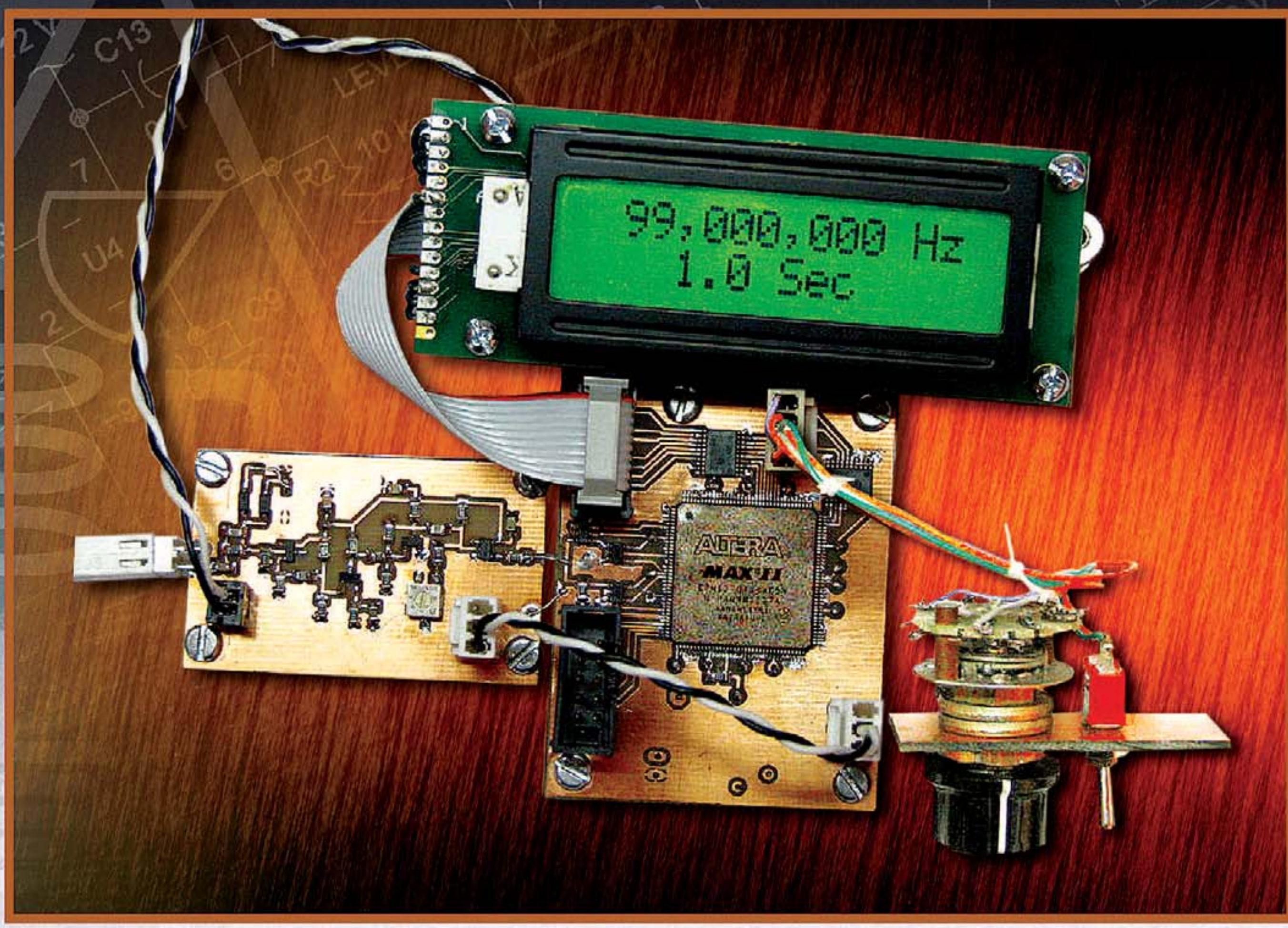
A Forum for Communications Experimenters

Issue No. 279



AV7VM designed a frequency counter using a complex programmable logic device (CPLD). The project was an opportunity to learn about this new class of ICs. Here is the resulting counter, ready for packaging.

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About the Cover

Gary Richardson, AV7VM, wanted a better frequency counter than a previous project. He thought about using a field programmable gate array (FPGA) but realized that a complex programmable logic device (CPLD) would be better suited to the task. The project was an opportunity to learn about this new (to him) class of ICs, so he began reading about them. The resulting frequency counter is described in his article, "Another Frequency Counter for the Experimenter."

In This Issue

Features

Index of Advertisers

Larry Wolfgang, WR1B

Empirical Outlook

Advancing Technology

I rarely use this space to write about articles you will find inside these pages. I am making an exception this time because there are several articles that seem to have caught my imagination. Perhaps they will catch yours, too!

Gary Richardson, AA7VM, describes a frequency counter that he designed. While the title may not sound that exciting (Another Frequency Counter for the Experimenter) there is one aspect of this project that had me scrambling to learn more. I have barely begun to grasp the concept of a field programmable gate array (FPGA) and how we can use these devices in Amateur Radio projects. Gary rejected the FPGA as being overkill, and instead decided to try a complex programmable logic device (CPLD).

A quick Internet search turned up quite a few websites with definitions and information about CPLDs. I learned that CLPDs can have thousands to tens of thousands of logic gates on the IC, while some FPGAs may have several million logic gates on a single array. It also seems significant that CLPDs have a non-volatile configuration memory, so you don't need a separate boot loader or configuration ROM to start the device. I imagine we'll be seeing more of these devices in *QEX* projects to come.

In another article, James Lee, N1DDK, describes a simple application for a micro-electrical-mechanical-system (MEMS) device with "Motion Based Electrical Power Control." This one really shocked me! Imagine an IC package that includes not only semiconductor technology, but also mechanical devices! A MEMS device is a tiny surface mount package that might include various sensors, actuators and transducers such as temperature, pressure and radiation sensors, along with inertial force (acceleration) sensors. There are even MEMS chips that include actuators such as microvalves for the control of liquids and gases, and optical switches and mirrors to redirect and modulate light sources. As if that isn't amazing enough, there are micropumps to create fluid pressures and microflaps to modulate airstreams on airfoils. Researchers have put these micro-actuators on the leading edge of airfoils of an aircraft and have been able to steer the aircraft using only these microminiaturized devices. You have to ask, "How do they do that?"

When you merge these miniature mechanical structures onto a common silicon substrate with various integrated circuit components such as bipolar, CMOS and other transistor and logic gate arrays, you are creating some truly versatile devices!

Two of the more interesting articles I found about MEMS devices are at **www.memsnet.org/mems/what_is.html** and **www.allaboutmems.com/memstechnology.html**. It might be a stretch to imagine ways we could use more of these devices in Amateur Radio applications, but I believe that some of you will do just that eventually. I have no doubt that some of our readers are even involved with the design and manufacture of MEMS devices. This may not seem all that amazing to them, but to many of us it will seem like pure science fiction. If you are involved with the design, manufacture or application of these devices professionally, we would love to hear from you.

There is another article in this issue that should have a more immediate direct impact on the design and construction of Amateur Radio equipment. Colin Horrabin, G3SBI, tells us about his work on the receiver front end of "The HF7070 HF/LF Communications Receiver Prototype," a high performance receiver that has been prototyped but is not going to find its way into manufacture. This receiver has some outstanding performance measurements. Colin's design uses a pair of H-mode mixers in an up converting, double conversion front end.

This may not seem like groundbreaking work. Colin developed the H-mode mixer in the early 1990s. We have been building crystal filters for many decades. The total of this design, developed by Colin, refined by several other experimenters for homebrew and commercial products as it results for the HF7070 prototype receiver provides some outstanding performance measurements, though. Without stealing all the thunder from Colin's article, when you have a 3rd order IMD Dynamic Range measurement of near 115 dB at 20 kHz signal spacing, you have a pretty amazing receiver! Imagine the receivers we could have if other commercial manufacturers started with this design and then further refined and developed it. (QST Product Review measurements of today's best receivers give 3rd order IMD Dynamic Range measurements in the range of 100 to 110 dB.) How far can we take the state of the art in receiver performance?

You Respond With Input

In the May/June issue I asked you for some input with regard to suppliers and advertising, and also about your thoughts on print versus digital publication of *QEX*. I received many thoughtful replies to those questions. Some offered suggestions for possible advertisers, and they seem worth a contact from our Advertising Sales staff. Thank you.

There were also a range of e-mails about things to consider with regard to digital publications. Few were in favor of a digital-only publication. Many of those who were interested in the availability of a digital version indicated a strong preference for a PDF-based format.

I want to assure you that no decisions are being made at this time. I will share your input as any discussion goes forward.

By now I hope you are anxious to dive into the pages of this issue, so go ahead. Enjoy!

Gary Richardson, AA7VM

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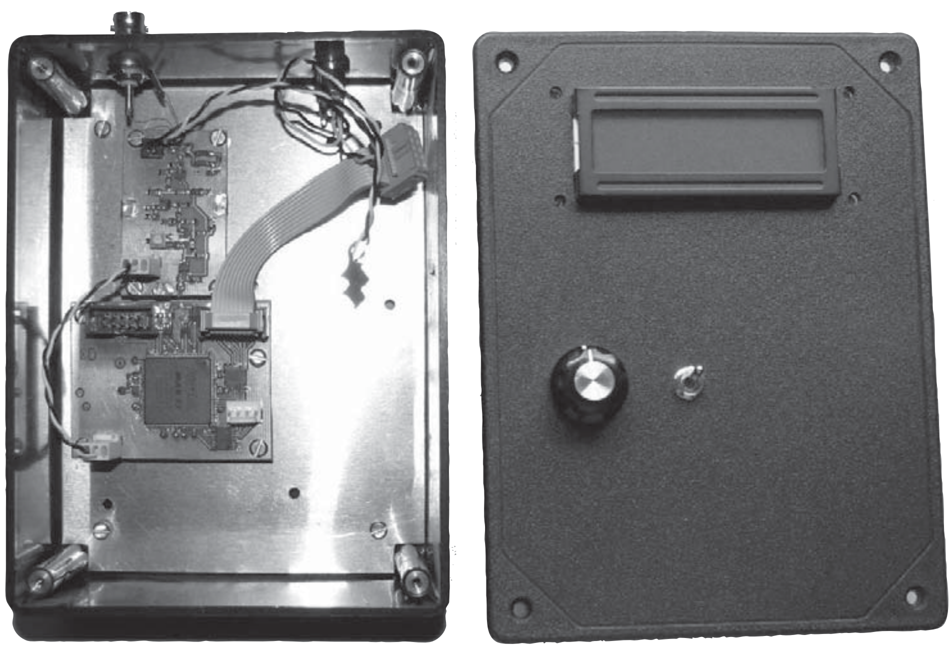
Another Frequency Counter for the Experimenter

A different approach to building an accurate frequency counter using a Complex Programmable Logic Device.



In 2011 I built a version of the frequency counter that Rubens Fernandes described in his May/June 2010 *QEX* article.[1] It didn't turn out too well because I used a relatively slow, low-power microprocessor. I then started thinking about how I might make a better counter. I had been very impressed by James Ahlstrom's articles describing his experience with FPGAs (Field Programmable Gate Arrays) and I thought that approach might work well for a counter.[2, 3] An FPGA, I soon discovered, would certainly be overkill, but a Complex Programmable Logic Device (CPLD) might do the job nicely

In simple terms a CPLD consists of a large number of configurable logic elements (LE). Each LE contains a section called a look up table (LUT) that can perform any logic operation on four variables. An LE can perform single-bit addition or subtraction and groups of LEs can be configured to perform multi-bit addition or subtraction. An LE also contains a register that can be configured for D, T, J-K or S-R flip-flop operation and groups of LEs can be configured as multi-bit registers. Unlike an FPGA a CPLD has no random access memory. Notes 2 and 3 point toward good discussions of FPGAs, much of which is applicable to CPLDs.

A CPLD is quite different from all other types of hardware that I had been exposed to. I knew nothing about configurable hardware devices or hardware description languages. So, like James Ahlstrom, I bought some books and began reading and searching the internet for information.[4, 5] That search soon brought me to *MyHDL*.[6] *MyHDL* is a *Python*-based tool that greatly simplifies the generation of HDL designs. To quote the *MyHDL* manual "*MyHDL* is a free, open-source package for using *Python* as a hardware description and verification language. *Python* is a very high level language, and hardware designers can use its full power to model and simulate their designs. Moreover, *MyHDL* can convert a design to *Verilog* or *VHDL*." For me, *MyHDL* was a lifesaver; I doubt if I would have begun this project otherwise. Figures 1 and 2 illustrate a few of the features of *MyHDL*.

[1]Notes appear on page 6.

Dissecting the Code

Figure 1 is the *Python* code for the top-level module of the counter (*counterMain*) and one utility module (*syncSig*). The arguments to *counterMain* are the names of the signals assigned to I/O pins of the CPLD. The variables *mode*, *reset*, *interval* and *intervalRaw* define signals required by these modules. A total of five modules are instantiated (created): Three of *syncSig* (S1, S2, S3) one of *counterMainP* (CMP, the main processing module) and the *counterMain*, the logic that drives the *indicator* signal and inverts the *intervalRaw* signal. The *counterMainP* module instantiates modules for counting and display processing and defines the logic necessary to initiate a counting cycle and display the results. The *syncSig* modules insure that the *mode*, *intervalRaw* and *reset* signals are updated synchronously with the values of the asynchronous input signals.

Figure 2 is a small module that illustrates how a finite state machine (FSM) is defined with *MyHDL*. An if-elif-else structure involving the state variable is recognized and converted to *Verilog* or *VHDL* case statements. The states are defined in a *Python enum* statement. In the third line of this code the *state* signal is defined and given an initial value of STATE1. The logic specified in the *clockDivide* function is updated on every negative edge of CLK_50. The FSM remains in STATE1 for three clock periods and in STATE2 for two. The *clockDivide* module is one I used during testing to generate a 10 MHz clock from the 50 MHz MAX II clock signal on the development board I used.

Getting Started

While there are a number of suppliers of CPLDs the choice of which one to use was easy — Altera, the only one I could find that provided free, adequate development software.[7] The Altera Quartus II package provides everything necessary to generate the configuration data, including a device programmer. There is also a limited version of a commercial simulation program available that operates on the HDL code and is much faster than the *MyHDL* simulator, but is more difficult to use. It has a waveform viewer that I found useful at times.

A Terasic MAX II Micro kit was very helpful in getting started.[8] The MAX II board has on it Altera's largest CPLD, four pushbuttons, eight LEDs and a 50 MHz oscillator for the clock. It also has two prototyping

```
def syncSig(clk, pIn, pOut, nbits):
    p1 = Signal(intbv(INACTIVE_HIGH)[nbits:])
    @always(clk.negedge)
    def logic():
        p1.next = pIn
        pOut.next = p1
    return logic


def counterMain(clk, resetIn, counterSig, externalSig, modeIn, intervalIn, gate,
                sigSelect, lcdBus, E, RS, RW, indicator):

    mode, reset = [Signal(bool(INACTIVE_HIGH)) for k in range(2)]
    interval, intervalRaw = [Signal(intbv(0)[2:]) for k in range(2)]

    S1 = syncSig(clk, modeIn, mode, 1)
    S2 = syncSig(clk, intervalIn, intervalRaw, 2)
    S3 = syncSig(clk, resetIn, reset, 1)
    CMP = counterMainP(clk, reset, counterSig, externalSig, mode, interval, gate, sigSelect,
   lcdBus, E, RS, RW)

    @always(clk.negedge)
    def logic():
        indicator.next = ~(gate[0] | gate[1])   # control for LED

        # invert the bits. The external circuit produces values 3, 2, 1
        # whereas what is needed is 0, 1, 2
        interval.next = ~intervalRaw

    return instances()
```

Figure 1 — The *Python* code for the top-level module of the counter (*counterMain*) and one utility module (*syncSig*).

areas that provide access to a large number of the CPLD I/O pins. Configuration data can be downloaded from a computer via a USB cable, which also supplies power to the device. The MAX II can also download configuration data to a user board so a separate USB blaster cable is not necessary. The *Verilog* code for several demo projects is also provided. It was somewhat instructive to examine that code but I found most of it to be incomprehensible due to an almost total lack of comments and my ignorance of *Verilog*.

The counter can measure frequencies of just less than 100 MHz. The counter could perhaps handle higher frequencies but I allocated only enough resources (registers) to count to and display an 8-digit value. Frequency can be measured over a 0.1, 1.0 or 10.0 second interval. Period is measured by counting cycles of the clock signal for 1, 10 or 100 cycles of the input signal and the result expressed in microseconds. A 10 MHz TCXO provides the timing reference for the measurements and the clock for the CPLD.

Counter Hardware

Figure 3 is a photograph of the counter hardware. The CPLD, and five ICs are mounted on the top side of the primary board. The TCXO, voltage regulator and a few discrete parts and bypass capacitors are on the bottom. The three chips to the left of the CPLD (barely visible) are a switch IC and two gates. The switch connects the appropriate signal (the external signal or clock, depending upon the mode) to an input pin routed to the counting logic. The clock is connected to another input pin that is routed to the period timing logic. Both of these signals are gated. The gates are enabled or disabled by signals from the CPLD as required by the mode and measurement interval. I had originally thought that control of the counting process could have been done within the CPLD but that proved to be impractical. The two ICs near the upper-right corner of the board are '541 octal buffer/drivers. They buffer the signals going to the display and the mode and interval inputs. Perhaps they were not absolutely necessary, but were used to buffer similar signals on the MAX II board, so I used them also. I used wires to connect power to several areas on the bottom of the board to minimize breakup of the ground plane. The smaller board on the left is an amplifier and a Schmitt trigger output stage. That circuit is the same as the one designed by Rubens for his counter. The display is a 2 × 16 LCD operated in the 4-bit mode.

I had planned to have a board for the CPLD made by a commercial company but was unsuccessful in uploading my Gerber files. So before looking for another board manufacturer I decided to attempt making a

```
def clockDivider(CLK_50, clk10):
    # Divide 50 MHz clock by 5
    State = enum('STATE1', 'STATE2')
    state = Signal(State.STATE1)
    k = Signal(intbv(0, min=0, max=3))

    @always(CLK_50.negedge)
    def logic():
        if state == State.STATE1:
            if k == 0:
                k.next = 1
                clk10.next = not clk10
                state.next = State.STATE2
            else:
                k.next = k - 1
        elif state == State.STATE2:
            if k == 0:
                k.next = 2
                clk10.next = not clk10
                state.next = State.STATE1
            else:
                k.next = k - 1
        else:
            state.next = State.STATE1
    return logic
```

Figure 2 — Here is a small module that illustrates how a finite state machine (FSM) is defined with *MyHDL*.

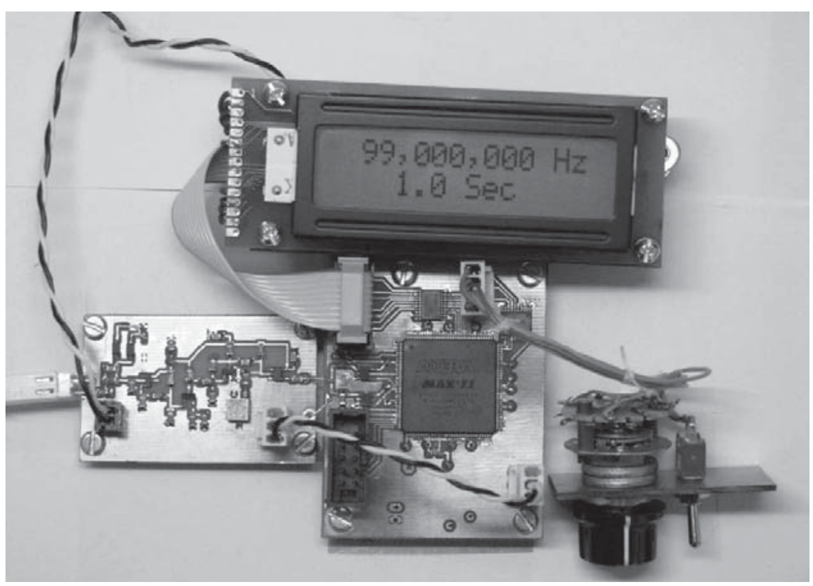


Figure 3 — This photo shows the basic counter hardware.

board myself. Much to my surprise it turned out pretty well. I used Toner Transfer Paper to generate the masks.[9]

This design required 95% of the EPM1270T144 resources. In terms of lines of *Verilog* code, about 20% was dedicated to the count processing and the rest to the display processing. The display processing would have been much simpler in a microprocessor but my goal was to do everything in the CPLD.

Conclusion

I don't have access to a GPS or Rubidium referenced signal generator so I can't tell exactly how accurate the counter is. The measurement of the 10 MHz output of an AADE Precision Frequency Reference is within 1 or 2 counts of 10 MHz. The output at 10 MHz of a signal generator I have is about 3 Hz lower than the 10 MHz AADE signal. To get the 99 MHz shown in Figure 3 the signal generator had to be set about 30 Hz higher. While this doesn't say anything about accuracy it does imply that the counter is working reasonably well at this frequency. Measurement of the AADE period at 0.1, 1, 10, 100 kHz and 1 MHz is exact.

This has been a very interesting project and has kept me occupied for quite a while. It provided the opportunity to learn a bit about a few new things, though I'm sure I've only scratched the surface of hardware design with configurable devices. I thank Rubens Fernandes and James Ahlstrom for their very interesting *QEX* articles.

The files for this project are available on the *QEX* files website at **www.arrl.org/qex-files**. Look for the file **5x13_Richardson.zip**. The file includes the Python logic design files, and the Eagle schematic circuit board layout files.[10]

Notes

[1]Rubens R. Fernandes, "A Frequency Counter for the Experimenter", *QEX,* May/June 2010, pp 10-15.

[2]James C. Ahlstrom, "An All-Digital Transceiver for HF", *QEX,* Jan/Feb 2010, pp 3-8.

[3]James C. Ahlstrom, "An All-Digital SSB Exciter for HF", *QEX,* May/June 2008, pp 3-10.

[4]Bob Zeidman, *Designing with FPGAs & CPLDs*, CMP Books, 2002.

[5]Donald Thomas and Phillip Moorby, *The* Verilog *Hardware Description Language*, Fourth Edition, Kluwer Academic Publishers, 1998.

[6]You can find more information about *MyHDL* and download the package at **www.myhdl.org**.

[7]Information about Altera CPLDs and the development software are available at **www.altera.com**.

[8]Altera has formed a partnership with Terasic as a global supplier of Altera development kits. For information about the MAX II Micro kit, go to **www.terasic.com** and search for MAX II.

[9]The Toner Transfer Paper system for creating circuit board masks is available from Digi-Key (**www.digikey.com**). There is more detailed information about this material on the Pulsar website at **www.pulsarprofx.com**.

[10]The various files for this project are available for download from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **7x13_Richardson.zip**.

Gary Richardson, AA7VM, was first licensed as KN5WHO in 1957. Interest in Amateur Radio waned in subsequent years from the pressures of school and work. Gary earned an MSEE degree from Michigan State University in 1967 and spent much of his career designing software for embedded microprocessors in medical systems. He was licensed as AA7VM in 1994.

Hamish Kellock, OH2GAQ

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A Microwave Transverter Controller

Flexible, sophisticated control of multiple transverters from a single point.

For many years I have been interested in VHF, UHF and microwave experimentation, and this controller is one of the outcomes of these interests. The unit is self-contained, and collects many of the functions needed to operate and control a microwave station on many bands from 1296 MHz upwards. Incorporated in the design is a formalized and relatively simple interface to connect the controller to any high or low power microwave transverter (such as 1296 MHz, 10 GHz, 24 GHz and higher frequencies), so that full multi-band station control can be made from a single unit, and band selection can be made also, with direct readout of the transmit frequency and/or receive frequency. Up to four different transverters can be connected at any one time to the controller.

A conventional way to proceed to make a microwave station, especially for portable operating, is to use a small commercial transceiver (such as an ICOM IC-706 or similar) at 144 MHz, then use it to drive the microwave transverter(s). For fixed station use a similar approach is used, but with perhaps a higher quality transceiver being employed. A collection of additional control and sequencing units, interfaces to computers, frequency stabilizing units and so on are employed to complete the station. As most commercial transceivers with acceptable performance are generally able to deliver up to 100 W PEP SSB, the electrical efficiency is somewhat low at the level needed for a typical transverter, which is generally 1 to 10 mW, or maybe up to a few watts if the transverter has an appropriate attenuator either built in or used externally.

Figure 1 shows the block diagram of the OH2GAQ Microwave Transverter Controller, while Figures 2, 3 and 4 show photos of the finished prototype. The overall unit size is 260 mm wide × 95 mm high × 315 mm deep, which is quite comparable to a typical table-top transceiver. The power supply for the controller is 12 V and 26 V dc.

The functions included in the Controller are:

- Microphone preamp and simple audio clipping, level control and band-pass filtering.
- Transformer-isolated input from a computer-generated analog audio signal.
- Switch selected audio source.
- Upper/Lower SSB generation (at 9 MHz, filter type) and up-conversion to an output signal in the range of 28 to 30 MHz.
- A direct digital frequency synthesizer to generate the LO for up-conversion of the 9 MHz SSB signal. The DDS reference is derived from a 10 MHz Rubidium reference oscillator incorporated in the unit.
- A Rubidium reference oscillator and low-noise distribution amplifier, including a 50 Ω line driver to route the reference to a bank of remotely located microwave transverters.
- A 144 MHz transverter, with 10 mW output, using the same reference oscillator at 116 MHz as the DDS to ensure adequate frequency stability. The transmit side is supplied with 28 MHz SSB; the receive side is fed to an external receiver.
- An ICOM compatible CI/V output to allow control of a typical ICOM receiver such as the R75, used by the author as a tuneable IF at 28 MHz.
- Muting control for the receiver during transmit.
- Receiver audio conditioning and an isolation transformer to allow connection of the received signal to a computer analog audio interface.
- Control input from a computer to allow computer-controlled receive/transmit switching when this mode is selected.
- A microprocessor-based control module to control and monitor all of the above, and with interfaces to a computer (using either R232 or USB-2) and up to four microwave transverters. The transverter interface has a transmit signal output, a transmitter health signal input and a couple of analog monitoring signal inputs as well.

Implementation Details

The main functional blocks in the controller are made from a combination of commercially available units or kits, possibly with some modifications, and some items that are designed as part of the controller. The commercial items used have generally been chosen on the basis of performance versus cost and/or availability.

Modules that have been designed as part of the project have normally had schematic capture and layout done by using the *Eagle CAD* tool, although the SSB generator and crystal PLL were originally designed using a simple computer drawing tool. *Eagle* has been an excellent tool for schematic capture and later layout. The main controller board was manufactured by a circuit board manufacturer, because it has large numbers of through holes. All the others were made at home by using a laser printer onto photo paper, then thermally transferring the pattern onto the circuit board stock before etching. There have been several good descriptions of this method, which can generally be found by searching the Internet or looking at the ARRL or RSGB *Handbooks*. The most challenging board was the crystal PLL oscillator, where the PLL chip (ADF4112)

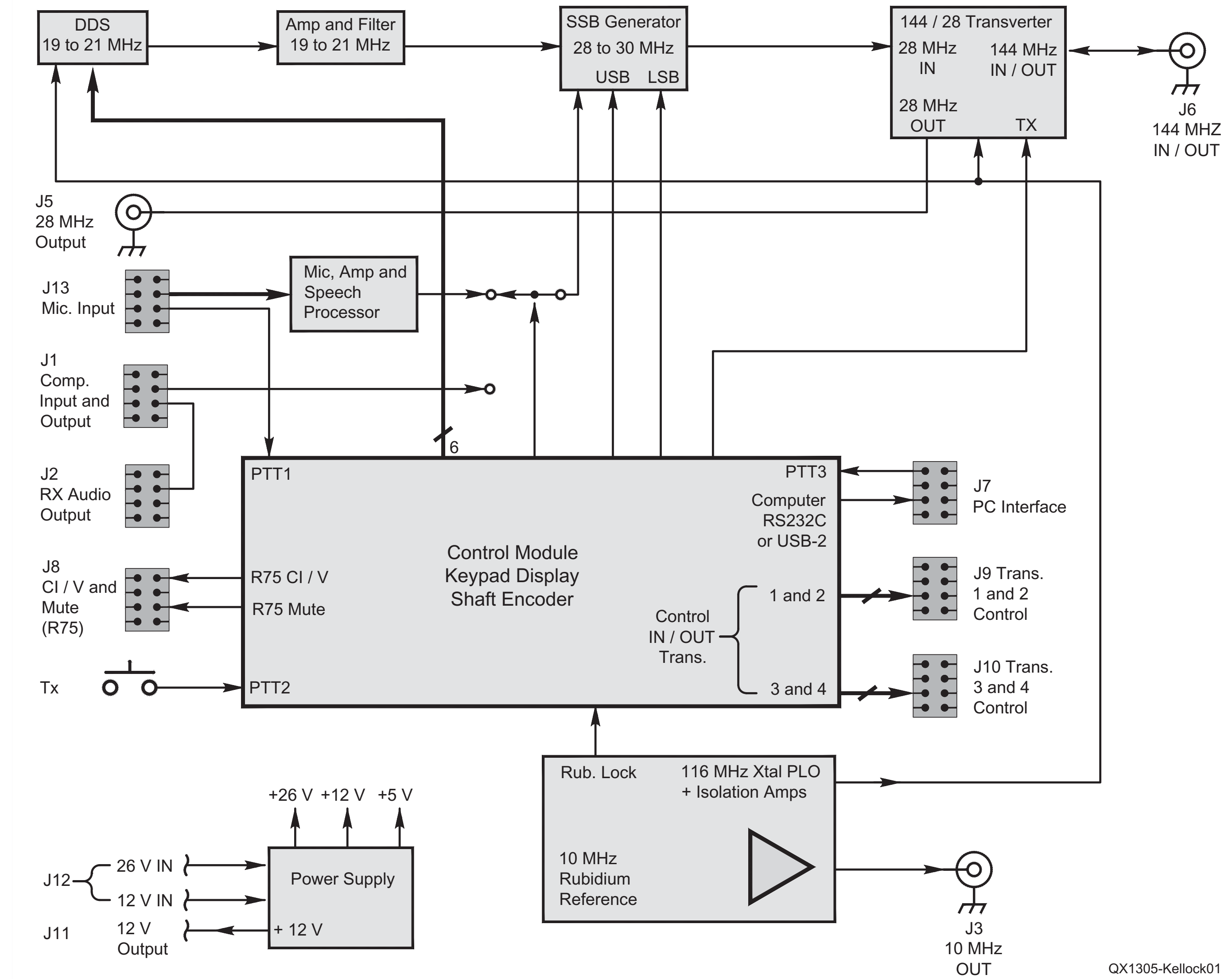


Figure 1 — Block diagram of the microwave transverter controller showing all the main functional blocks and external interface connectors.

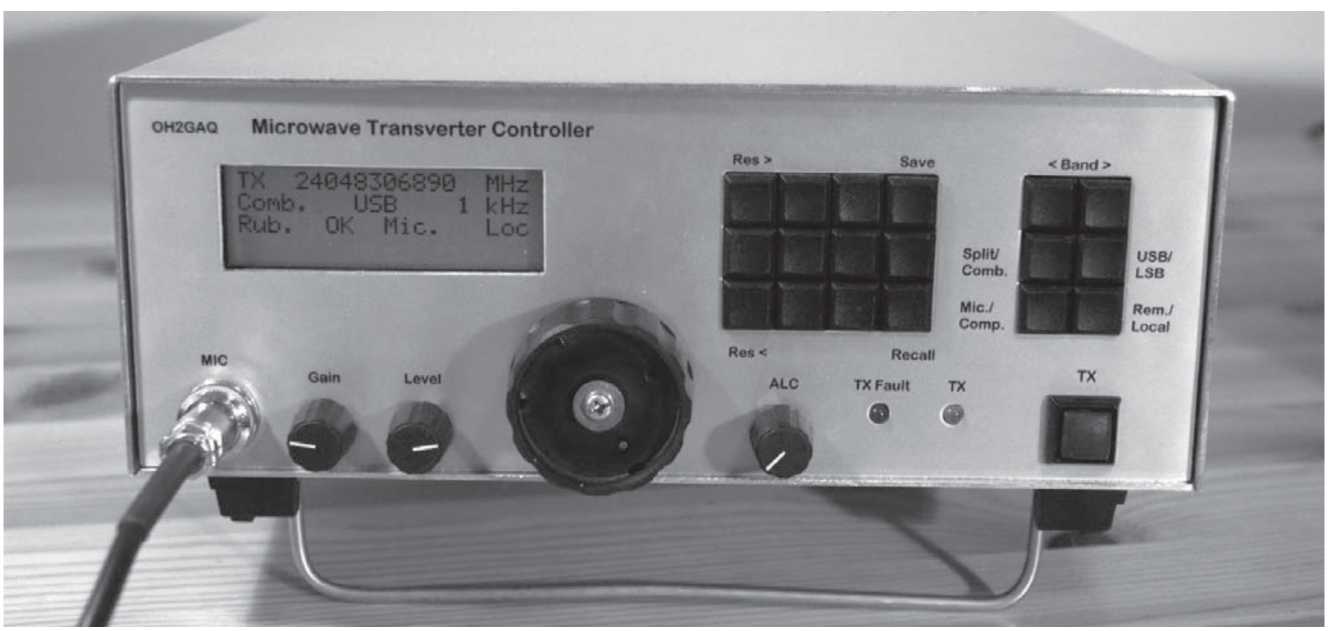


Figure 2 — An overall view of the microwave transverter controller.

was in one of those incredibly small TSSOP packages.

I used a Datum LPRO Rubidium frequency reference. In order to ensure that the reference signal was as clean as possible with reasonably low phase noise, the 24 to 26 V dc power supply should have very low ripple and noise. The external power supply I used is a linear supply with less than 6 mV of ripple and noise. Of course, a GPS disciplined crystal oscillator could also be used, but the Rb reference provides adequate stability, and was available at low cost.

The 10 MHz signal from the reference is passed on to a buffer amp using an LM7171 op amp to drive the remote transverters and a signal conditioner using a 74AC04 CMOS buffer and signal shaper. This provides the input to the crystal PLL oscillator, which in turn provides the internal 116 MHz reference for the DDS and 144 MHz transverter. A simple transformer is used at the input to provide impedance matching and a voltage step-up. Figure 5 shows the schematic of the signal conditioning following the LPRO oscillator. It includes the 116 MHz isolation amplifiers, which are in the same module. A photo of the distribution amplifier module is shown in Figure 6.

The DDS unit uses James (WA1FFL) Hagerty's Advanced Direct Digital VFO circuit board employing the AD9951 DDS chip. (See the Further Reading section at the end of this article.) The reference oscillator is replaced with the 116 MHz from the crystal PLL and the AD9951 is directly controlled by the main controller unit without using the microcontroller on Jim's board. Other functions are used as-is. The filters have been replaced with lower frequency cut-off units since the DDS has to only generate signals from 19 to 21 MHz. This board provided a quick and effective way to get the DDS functionality. The DDS has adequate performance for this application with broadband noise and general spurs being more than 60 dB down. There are a couple of DDS spurs that exceed this level. This board is followed by a band pass filter and amplifier using a 2N5109 transistor to increase the available output to drive the 9 to 28 MHz double balanced mixer in the SSB generator.

I used a Down East Microwave model 144-28INT kit as the 28 to 144 MHz transverter, with the major modification being to supply the 116 MHz LO from the external crystal PLL oscillator unit. The modification can be accommodated on the 144-28INT board quite simply, with only one short wire link being needed.

The 116 MHz crystal PLL (the same circuit with slightly different component values is used in all my crystal PLL applications in various transverters) uses

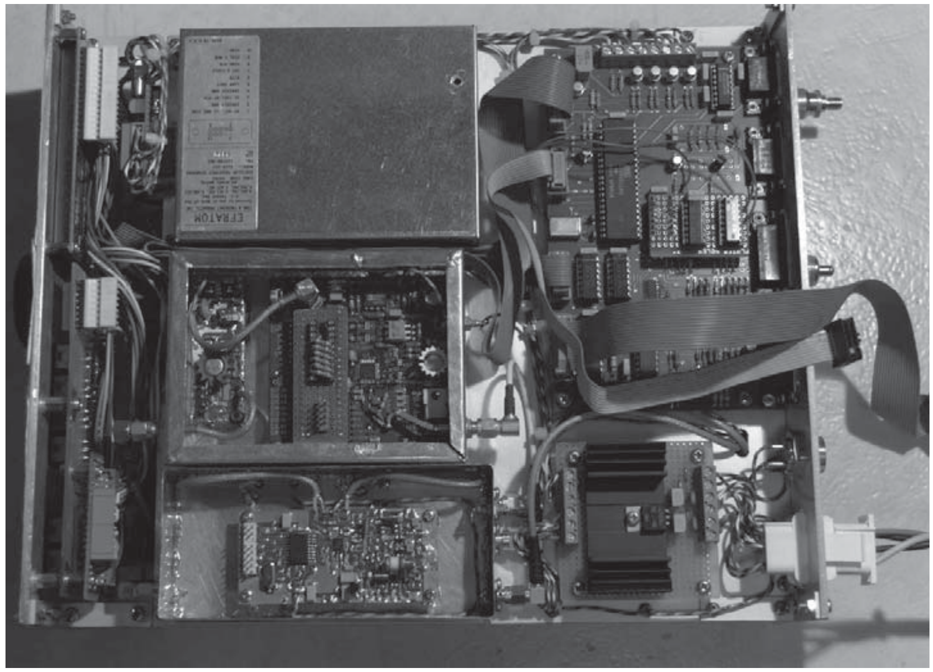


Figure 3 — Top view of the prototype controller, showing (clockwise from the top left) the 10 MHz Rubidium reference, the controller board with external interfaces, a small power supply, the 116 MHz PLL oscillator and distribution amplifiers and the DDS synthesizer and amplifier. Note that the small protoboard plugged into the main controller board allows programming the processor in-situ. It is not needed for normal operation. Some cables have been unplugged to better show the DDS unit.

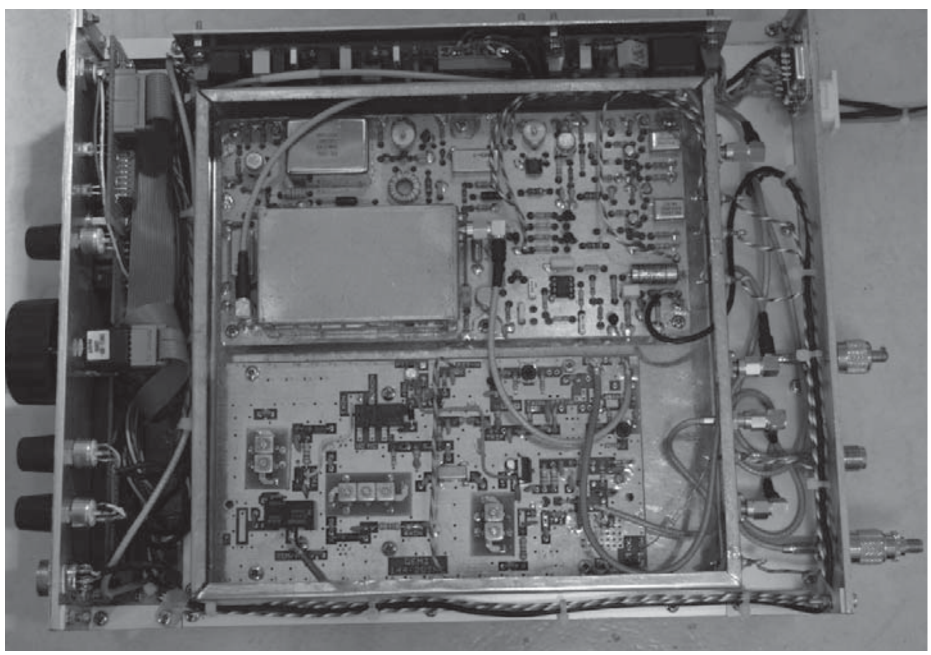


Figure 4 — Bottom view of the controller, starting from the bottom is the Down East Microwave transverter and the 28 MHz SSB generator, both in a tinplate enclosure. Above them is the audio processing amplifier and computer interface.

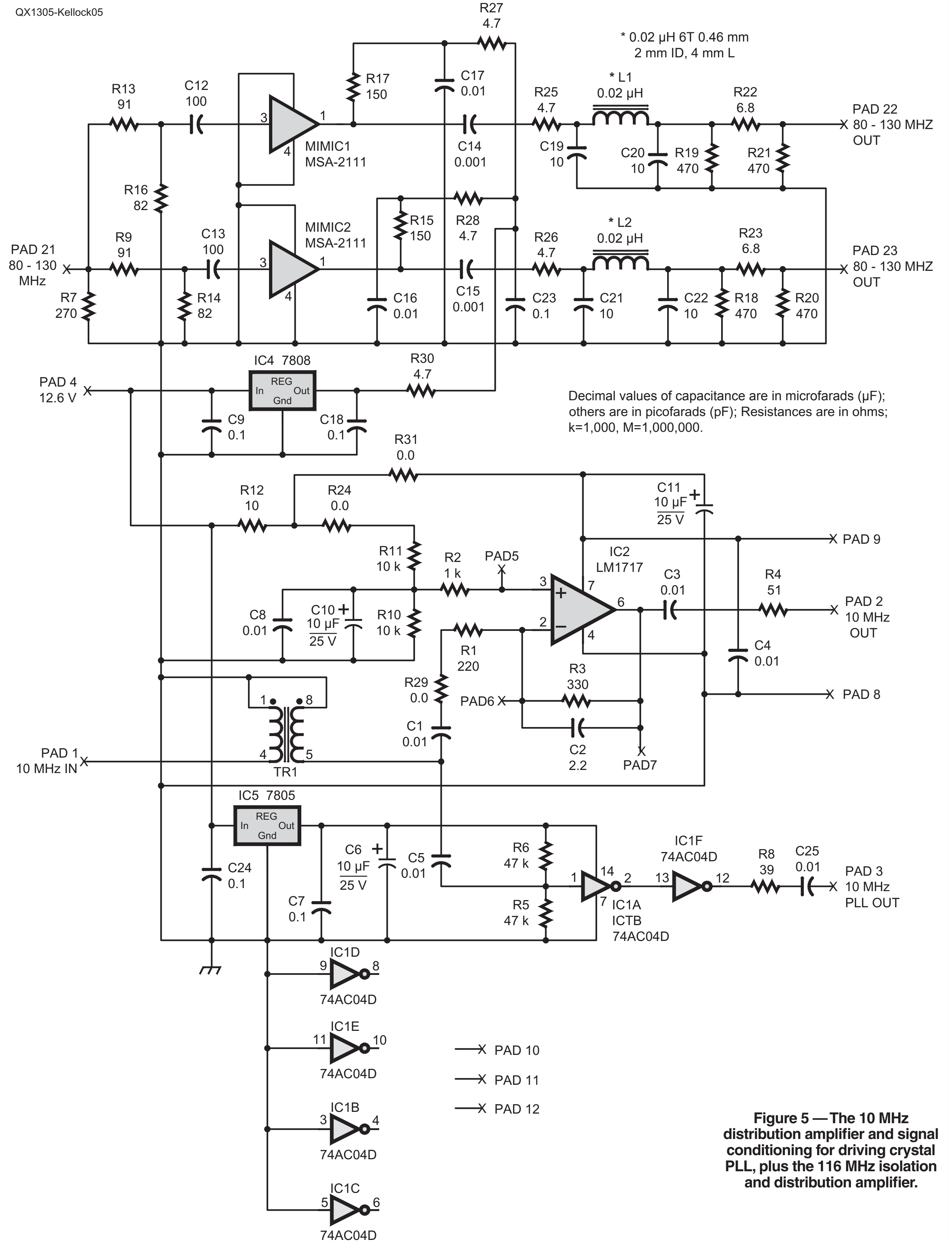


Figure 5 — The 10 MHz distribution amplifier and signal conditioning for driving crystal PLL, plus the 116 MHz isolation and distribution amplifier.

the stable and low-noise oscillator circuit developed by John Hazell, G8ACE, which seems to be similar to in approach to the work of John Stephensen, KD6OZH. To phase lock the crystal oscillator, a simple PLL using the ADF4112 has been added and a PIC 16F84A has been used to load the ADF4112 parameters at power up. A 40°C heater (Kuhne QH40A) is used to ensure the crystal temperature is constant. The circuit board is shown in Figure 7.

The circuit board has a solid ground plane on the back side; the crystal with its heater and one power regulator are mounted there. In order to reduce noise in the crystal oscillator, the ground plane on the component side is divided into two parts, with the digital ground being separated from the oscillator and PLL analog grounds in an attempt to reduce noise. There is a pin header on the right side of the board, which is used to program the PIC and later during operation to select which crystal frequency is being used (so a single program can be used for many applications such as LO for 1296 MHz, 2320 MHz, 24048 MHz or other bands). The software for the PIC is developed using Microchip's *MPLAB IDE*, where it is also possible to simulate the software operation and debug it, including the state of the input pins.

After the 116 MHz oscillator, a pair of simple resistive attenuators divides the 116 MHz signal, which is then amplified with a couple of HP MSA2111 MMICs to provide some isolation between the DDS and the 144/28 MHz transverter. The output level for each channel is about +8 dBm. These isolation amps have been combined with the 10 MHz reference distribution as mentioned earlier. The whole sub-system, including the 116 MHz PLL oscillator, is shielded in a small tinplate box.

Now let's have a look at the SSB generation. There are a few small circuit boards used to pre-condition the audio signal from a dynamic microphone. The signal is first amplified, filtered and clipped. Then, a relay is used to select either the microphone-derived audio, or that from a computer sound card (for JT65 or similar operating). Following this, the audio is passed on to the main SSB generator circuit board. Figure 8 is the schematic of the AF Preamp and switching scheme.

The main board for the SSB generator was designed and built many years ago and is a completely conventional 9 MHz filter type SSB generator. Two separate crystal oscillators generate 8.985 and 9.015 MHz carriers, one of which is applied to an SBL-1 mixer depending on whether USB or LSB is desired. The DSB output from the mixer is amplified, filtered and the resulting SSB is amplified before being mixed with the DDS-derived LO to up-convert the 9 MHz signal to 28 to 30 MHz. A band pass filter and amplifier follows the mixer. The resultant 28 to 30 MHz signal is applied to the Down East Microwave 28 to 144 MHz transverter. Figures 9 and 10 show the SSB generator schematic.

The overall performance of the SSB generation system is such that the carrier is suppressed by at least 60 dB and spurious outputs are at least 50 dB below the carrier level when measured at 144.5 MHz.

Miscellaneous analog signal functions like the computer interface are handled by a small unit. In order to keep noise down, isolated transformer coupling has been used between the computer soundcard (a Delta 44 card in my case) and the transverter controller.

The heart of the controller is the microprocessor based control board. A separate file, *kellock.zip*, downloadable from the *QEX* files website (**www.arrl.org/qexfiles**), contains schematics of the control

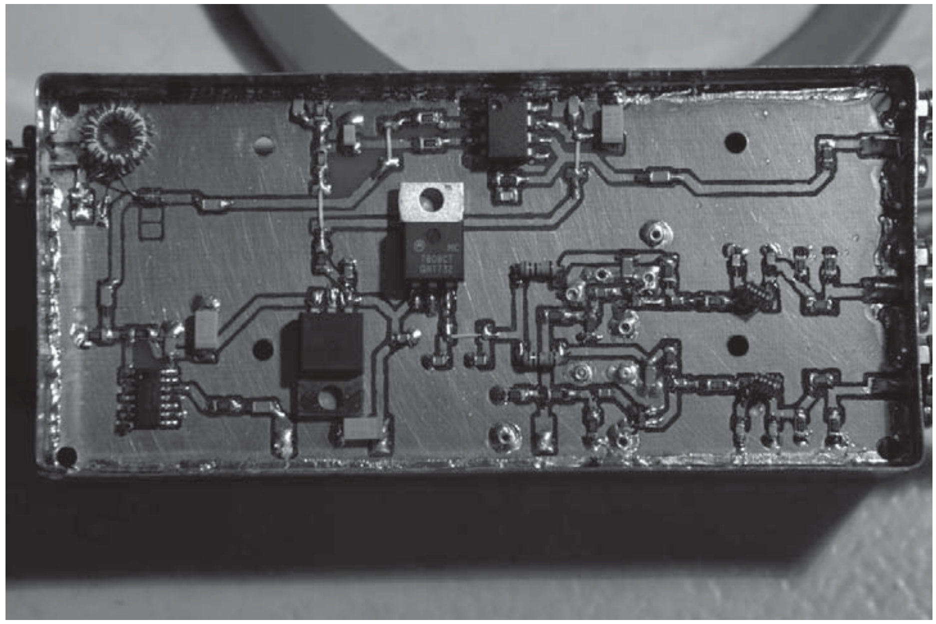


Figure 6 — The 10 MHz amplifier (top of module), 10 MHz interface for crystal PLL oscillator (lower left of module) and the two 116 MHz isolation amplifiers (lower right).

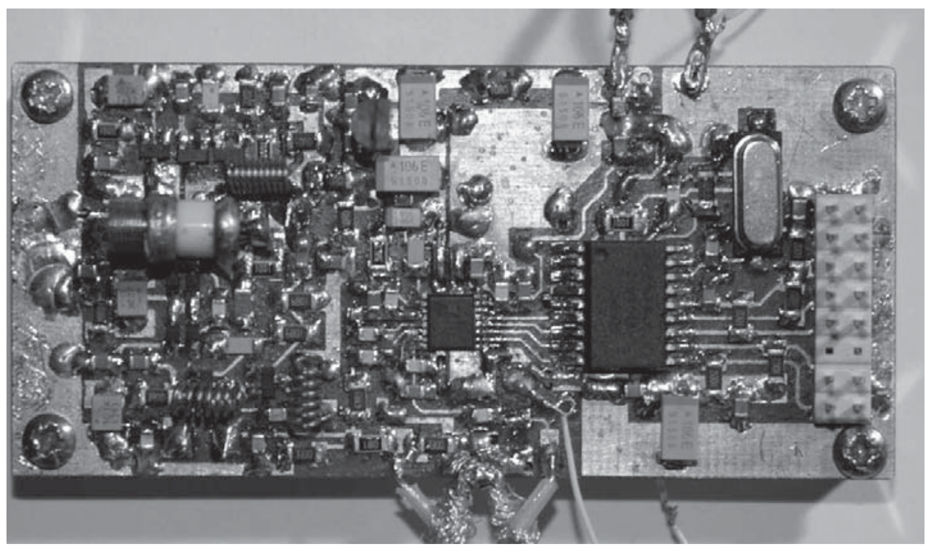


Figure 7 — The 116 MHz crystal PLL oscillator. The crystal and heater are mounted on the backside of the circuit board. The small HC-49 crystal seen here is for the PIC16F84A microcontroller (the largest chip on the board). The ADF4112 is the small chip almost in the middle of the board. Note the cut in the ground plane which extends under the ADF4112 chip. This separates the digital and analog ground planes.

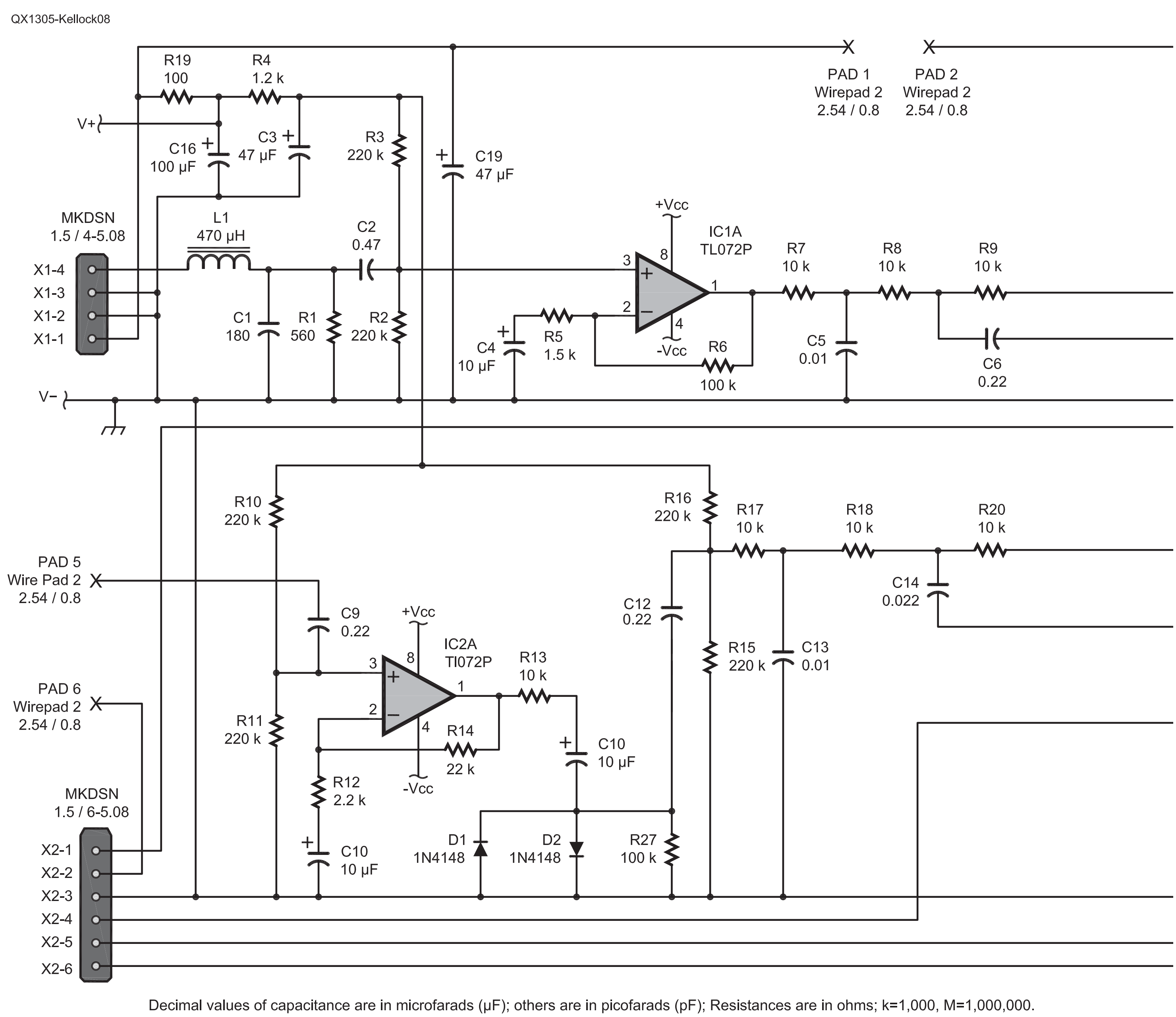
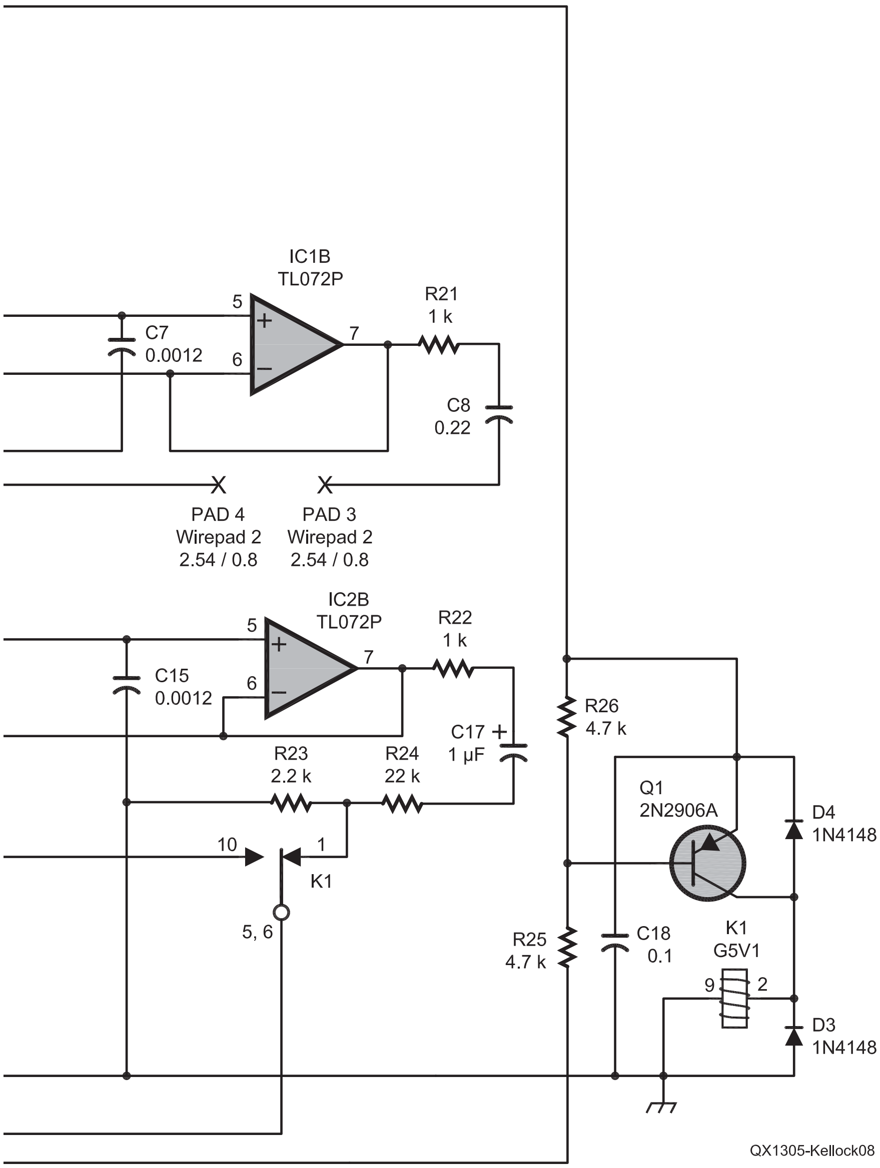


Figure 8 — Microphone pre-amp, audio clipping and local/remote switching for audio selection. P pads 1, 2 and 3, 4 and 5, 6 are used to place wire bridges, allowing a single sided circuit board. The microphone is connected to X1-4 and 12 V power to X1-1. Gain control (a 10 or 20 kΩ potentiometer) is connected to X2-1, X2-2 and X2-3. A second 10 kΩ potentiometer (for level adjustment) is connected to X2-5 and X2-3; the wiper is connected to the microphone input on the SSB generator in Figure 10.


IC1B
TL072P
5
6
7
R21
1 k
C7
0.0012
C8
0.22
PAD 4
Wirepad 2
2.54 / 0.8
PAD 3
Wirepad 2
2.54 / 0.8
IC2B
TL072P
R22
1 k
C15
0.0012
C17
1 µF
R23
2.2 k
R24
22 k
R26
4.7 k
Q1
2N2906A
D4
1N4148
10
1
K1
5, 6
R25
4.7 k
C18
0.1
K1
G5V1
9
2
D3
1N4148
QX1305-Kellock08

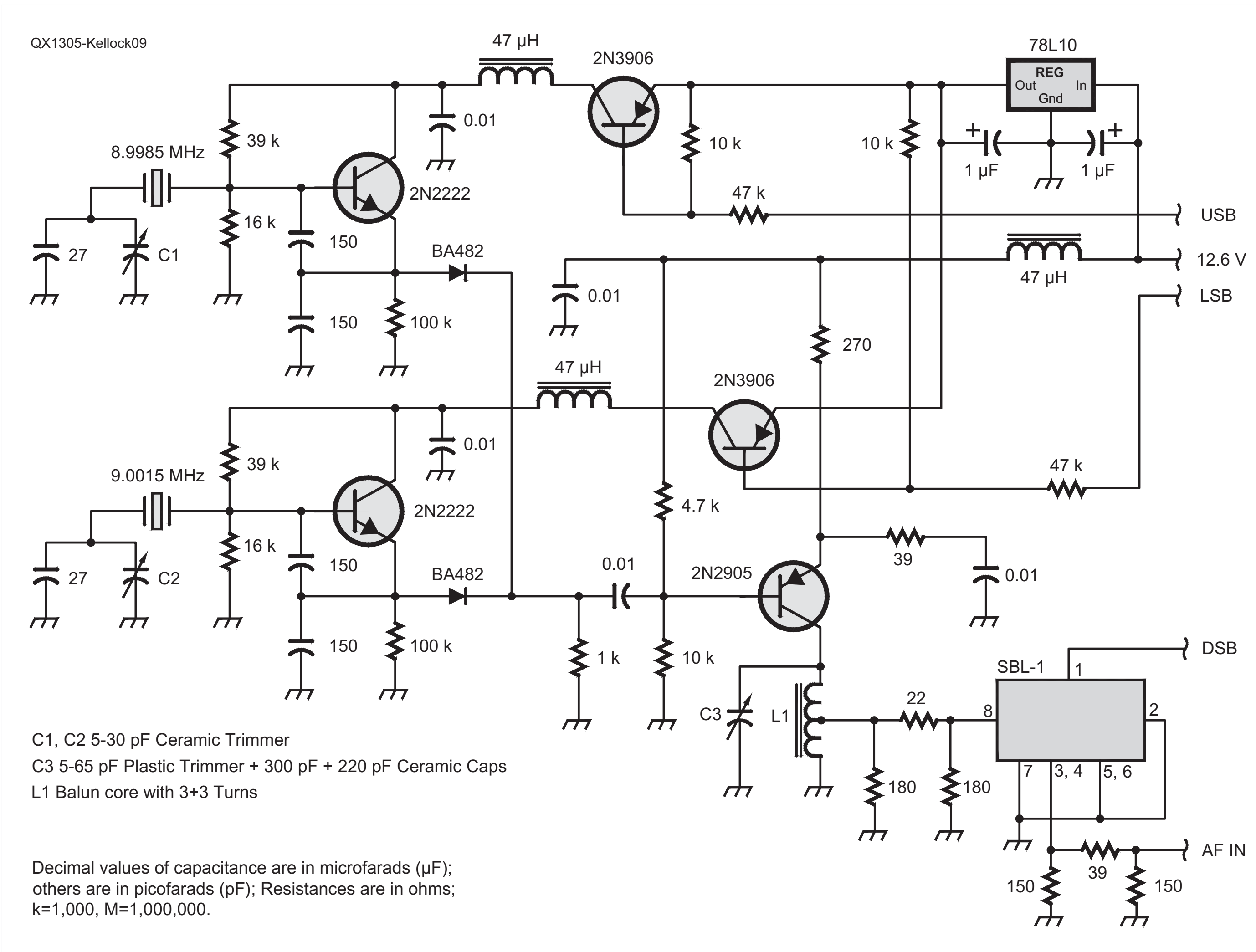


Figure 9 — The 9 MHz carrier oscillators, switching circuit and DSB modulator.

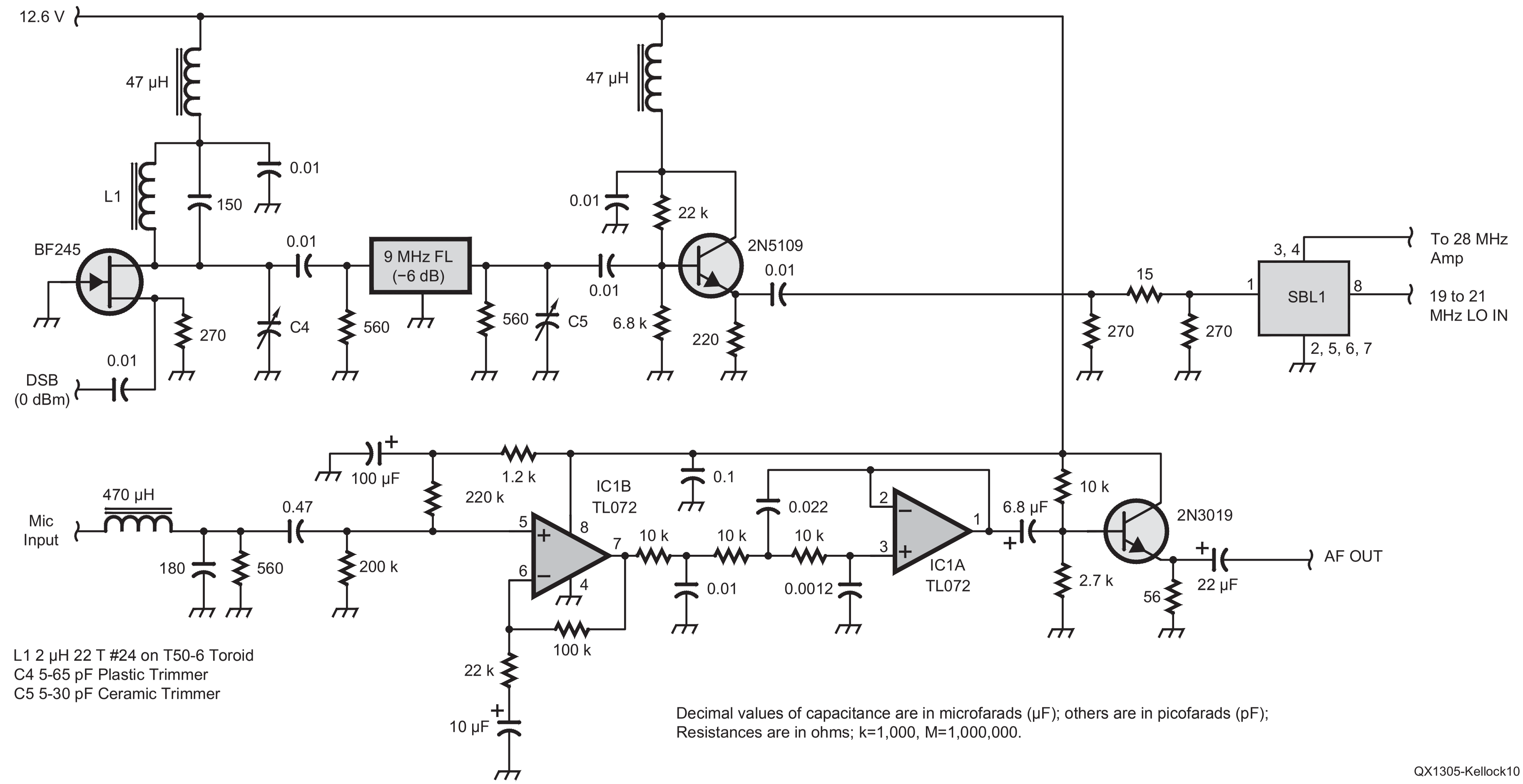


Figure 10 — The DSB/SSB filter and second mixer, along with the audio amplifier. The microphone input comes from the level control in Figure 8.

board and the PLL oscillator.[1] The processor used is one of the 8051 derivative processors, an Atmel AT89C51ED2. The circuit board was made for the 40 pin DIP variant of this processor, which is no longer generally available (other package versions are). The processor is augmented by adding a couple of MCP23S17 I/O extenders to give a few more input and output bits, plus an Analog Devices AD7888 8 channel A/D converter. The dial function is implemented by using a Bourns optical encoder, and IC9, IC8 and part of IC7 are used to implement an up/down direction sensing counter, which generates an interrupt to the processor for each count. The actual counting is done in the processor software. The CI/V interface (J8) is implemented using the remainder of IC7s inverters, plus a couple of transistors.

A partial RS232 or USB-2 interface (RXD, TXD and RTS) is implemented using either a MAX232 chip plus a DB9-F connector if RS232 is required, or an FTDI type DB9-USBD5-F module, which replaces the DB9-F connector on the circuit board. In that case, three wire jumpers replace the MAX232 chip, which is not required if a USB-2 interface is provided.

The interface to the external transverters is via J9 (two transverters) and J10 (two transverters). The digital outputs are driven by transistors, and the digital inputs are via transistors with resistor/diode clamping. The idea is that the actual transverters may be remotely located, and this scheme provides some protection and noise immunity for the digital inputs. The analog channels (two per transverter) are diode clamped. The outputs feeding them from each transverter should have a series resistance to limit the clamping current.

Transverter Interface Operation

Figure 11 shows the generic functions expected to be found in a transverter, which can be controlled by this unit. This is illustrated by the particular example of a 1296 MHz transverter.

First a word about the simple interface signals. I considered using a serial interface between the transverter controller and the transverter, however, I decided that the added complexity and potential reliability issues, versus the better monitoring and control that could be obtained, were not worth the complexity. The main disadvantage of the simple interface in practice is that fault causes are not explicitly shown remotely, and the analog monitoring of the forward and reflected power may be subject to a bit of noise if the cable run is particularly long.

[1]Notes appear on page 19.

The control signals to and from each transverter are:

- To transverter: Transmit Request.
- From transverter: Transmit OK

The transmit command is generated by the controller based on the band selected, and the PTT or remote transmit request. Following application of the transmit command, the controller waits for the transverter to signal TX OK. If this is not received within the specified time interval (set to 400 ms by default), the transmit command is de-activated and the transmit fault indicator is activated. Similarly, if a transmit fault condition occurs in a transverter during a commanded transmission, the same happens. In my transverters all the "fast" protection is self-contained in the transverter itself to ensure that any expensive amplifier transistors are protected and that the protection is not dependent on any remote signals.

The analog monitoring signals from each transverter are:

- From transverter: Forward power (analog voltage)
- From transverter: Reflected power (analog voltage)

The forward power is displayed as a bar graph reading on the bottom line of the 20 × 4 line display used in the controller. The scale factor can be individually set in software for each of the four transverters. Reflected power is also displayed.

Software Functions Implemented in the Controller

Figure 12 shows the display presently implemented in the controller, while Figure 13 shows the keyboard layout.

The software is implemented using 8051 assembler language compiled by the Systronix 8051 RAD51 IDE environment (which is available as a free download). Extensive use has been made of readily available math libraries for the 8051 (used in the frequency control of the DDS), plus some additional modules developed to handle 64 bit arithmetic. The software is downloaded to the processor flash over a serial line, using a small modification to the processor circuit board (a plug-in board and a couple of jumpers), with the *Atmel FLIP* (version 3.2.0) program.

User Interface Features

Band selection is done by simply incrementing/decrementing by each push of the **Band Up** and **Band Down** buttons. As there are only four bands it's quite quick. The readout is updated to give the full frequency readout of the selected band.

USB and LSB selection is toggled by successive depressions of the **USB/LSB** button. The current mode is shown on the display.

Local or remote control is selected by successive depressions of the **Remote/Local** button, and the current selection is shown on the display. Remote control is used with JT65, for example.

The receive and transmit frequencies can be separate or locked. The transmit frequency is controlled by the controller DDS. The receive frequency is controlled by the companion receiver, in my case an ICOM R75. This is the state in the "Split" mode. In the "Combined" mode, the controller queries the frequency set on the R75 through the CI/V interface, and sets the transmit frequency to the same value. Therefore, the tuning of the whole system can be done by the R75 tuning dial (and if the R75 frequency is not correct, there will be a small offset between receive and transmit). Successive depressions of the **Split/Comb** button toggle the mode.

The frequency increment represented by each unit of rotation of the tuning knob is selected from 1 Hz, 100 Hz, 1 kHz or 10 kHz. These are presented by successive depressions of the **Resolution Up** or **Resolution Down** buttons, with the presently selected resolution being shown on the display.

The current settings of USB/LSB, Local/Remote, Split/Combined, Tuning Resolution together with the Current Frequency (which also gives the band selected) can be stored into non-volatile memory with the **Save** button, and the last saved set can be recalled by depressing the **Recall** button

The state of the internal Rubidium standard is shown on the display. During the warm-up or if some other fault occurs, it is shown as either NOK or OK.

In addition to these features, the needed transmit and receive change-over functions are carried out by depressing the **Transmit** button on the panel, or by the remote control from the computer interface, if this is enabled. When in transmit mode, transmit frequency changes and USB/LSB mode changes are inhibited.

Operational Software Functions

The main functionality going on behind the scenes is the monitoring that's taking place in transmit mode. Any faults notified by the attached transverters cause immediate removal of the RF drive at 144 to 146 MHz and the removal of the transmit command from the selected transverter. An alarm condition LED (**TX Fault**) is illuminated. Scanning the KB, updating the display and LED status, updating the DDS frequency and so on also proceeds in the background.

There are many planned additions to the software, but if and when they will be

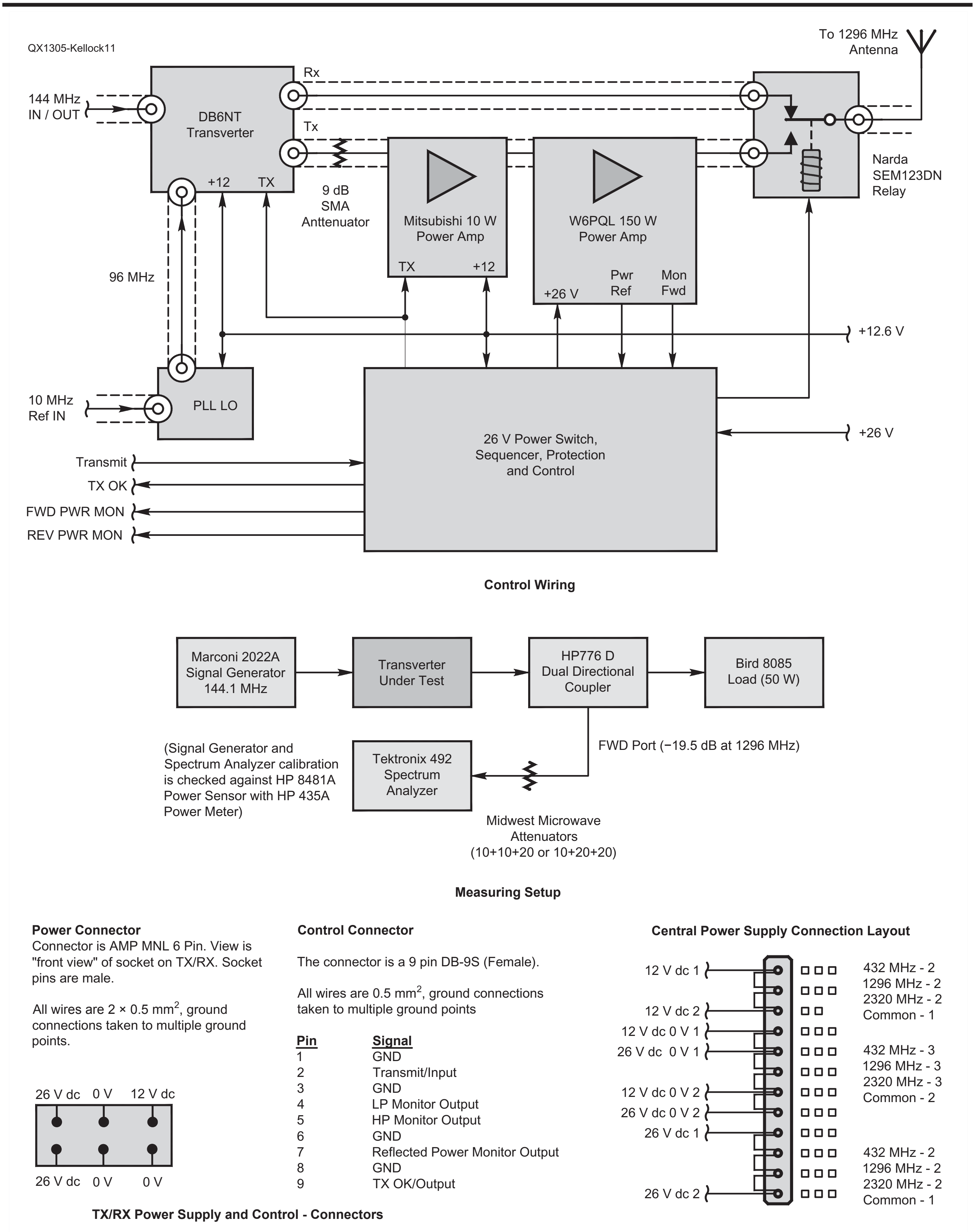


Figure 11 — A medium power microwave transverter for 1296 MHz, showing the main functional blocks and the control and monitoring signals.

implemented is another issue. The main future feature is implementation of a small subset of the CI/V commands on the controller so that a common computer can command both the R75 and MTC.

Figure 14 shows a picture of a subrack designed for three high-power transverters with the power supplies in the left hand side (26 V at 10 A and 12 V at 10 A) and fitted with a 1296 MHz transverter. The 1296 MHz transverter is mounted in a disused base station aluminum housing, which provides a good heat sink for the 150 W PEP power amplifier (built using one of Jim (W6PQL) Klitzing's 150 W kits). A DB6NT 1 W transverter drives a 10 W amp using a Mitsubishi MOSFET module, which then drives the 150 W amplifier.

Acknowledgements and References

There are many hams and others who have, generally unknowingly, contributed to the design and implementation of this system. The many excellent websites maintained by hams who wish to share their ideas and/or kits with others are too numerous to mention. More details of the implementation of some parts of the controller, such as the PLL oscillator, are contained on my website at **http://personal.inet.fi/private/oh2gaq/**. For those who are interested in more exact constructional details of some parts of the system, including *Eagle* design files or software source code, you can contact me at the e-mail address shown at the beginning of this article

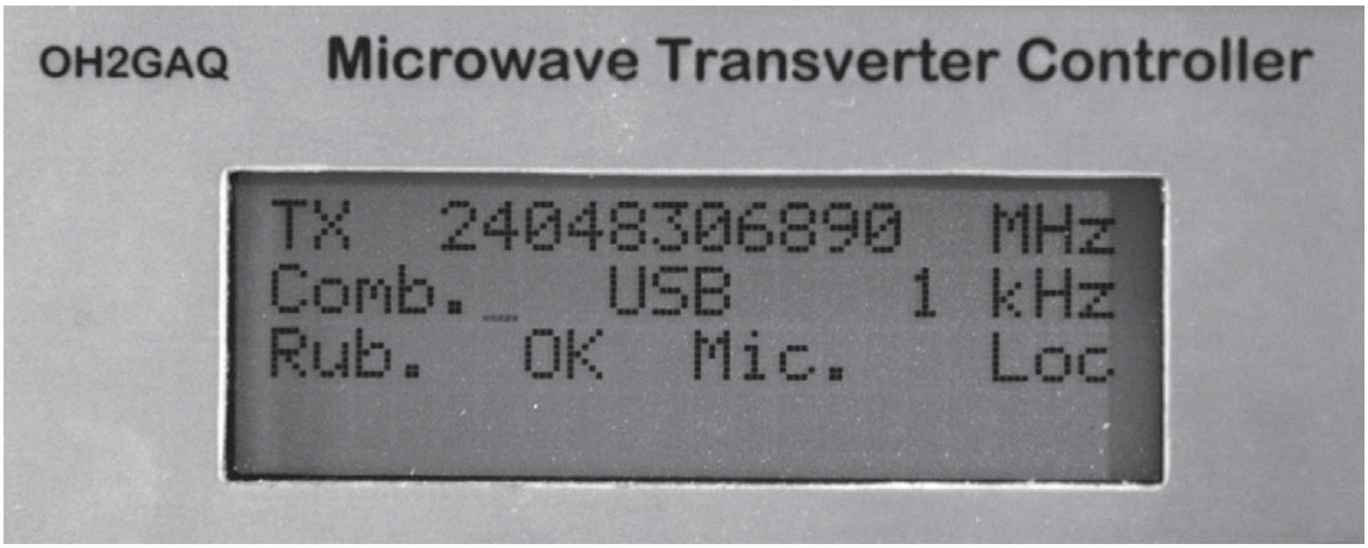


Figure 12 — Display of the MTC with present software. The top line shows the frequency, and implicitly the band in use (in this case 24 GHz). The mode is Combined (common receive and transmit frequency control). The transmit mode is USB. The transmit tuning resolution is 1 kHz (selected from 1 Hz, 100 Hz, 1 kHz and 10 kHz), but this is not relevant in Combined mode. The Rubidium Health is indicating OK. The microphone input is selected and the MTC is under Local control. The bottom line is used for the transmit output display. The 16 leftmost positions show forward power; the four rightmost positions show reflected power.

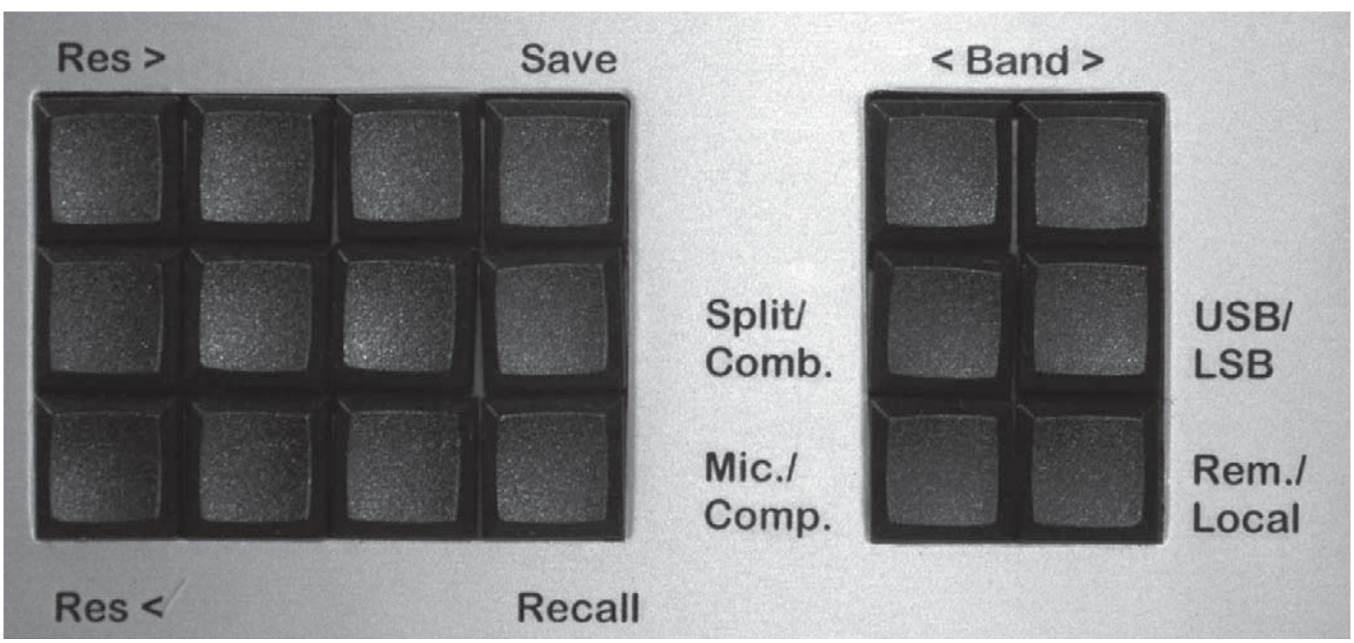


Figure 13 — The keyboard layout and key functionality with present software version. The main tuning functionality is controlled by the optically encoded main tuning knob when in Split mode

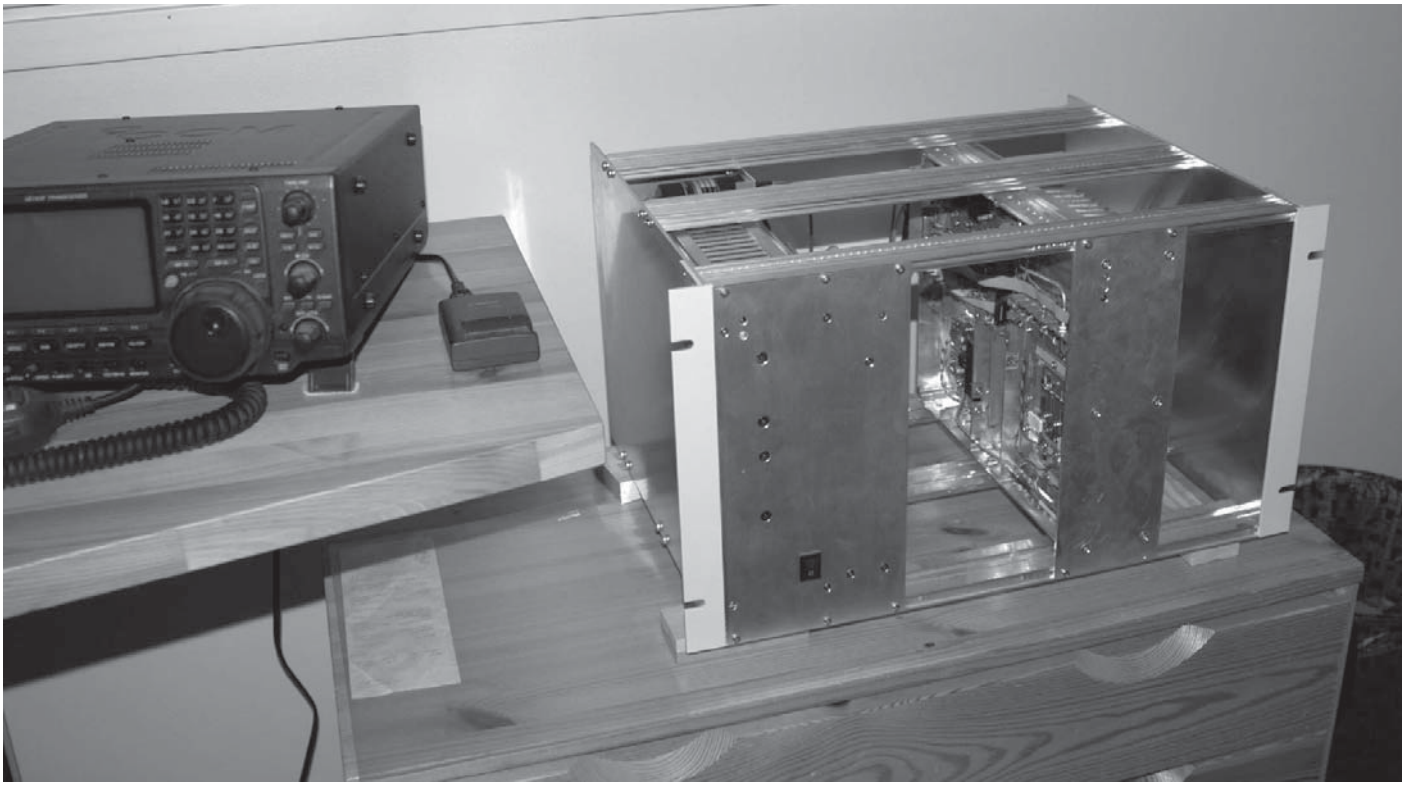


Figure 14 — Transverter subrack with the 1296 MHz transverter.

For Further Reading

I recommend the following articles and websites.

John Stephensen, KD6OZH, "A Stable, Low-Noise Crystal Oscillator for Microwave and Millimeter-Wave Transverters," *QEX*, Nov/Dec 1999.

John Hazel, G8ACE, "Constructional Notes for G8ACE MKII OCXO Sept 2010 V2," available from the G8ACE website at **www.microwaves.dsl.pipex.com/**.

The Analog Devices Data Sheet for ADF411x RF PLL Frequency Synthesizers.

The W6PQL website at **www.w6pql.com**. This site has several excellent articles covering microwave transverters and useful sub-systems, as well as actual kits for many items.

James D. Hagerty, WA1FFL "An Advanced Direct-Digital VFO," *QEX*, May/June 2008. See his website at **www.wa1ffl.com/** for DDS kits using the Analog Devices AD9951 DDS.

KO4BB's website at **www.ko4bb.com** has information about time and frequency control, measuring equipment and generally useful microwave related material.

KE5FX's website at **www.thegleam.com/ke5fx/** also offers time and frequency control information, measuring equipment and other useful microwave related material.

The Down East Microwave website at **http://downeastmicrowave.com/**.

The Kuhne Electronic website at **www.kuhne-electronic.de/en/home.html**.

Systronix RAD51 website at **www.systronix.com/RAD51/RAD51.htm**. This site details the Rapid Application Development Environment for 8051 family processors.

Note

[1]You can download a zip file of various files related to this article from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **7x13_Kellock.zip**.

Hamish Kellock, OH2GAQ, lives in Espoo, Finland. He has a Diploma in Applied Physics from the Royal Melbourne Institute of Technology (Melbourne, Victoria, Australia). He worked for several years in research associated with lasers and later with computer controlled measurements and instrumentation for RF measurements in Melbourne. For a short time he worked in the mineral processing industry, responsible for the development of computer controlled ore sorting equipment. In 1982 he moved to Finland to work in the telecommunications industry with Nokia. He held several positions over 28 years in the R&D area, covering network management, SDH transmission products, V5.2 multiplexers, microwave radio links and IP DSLAM products. He is now retired. Hamish has published several articles covering work with lasers, and holds patents in the mineral processing field as well as telecommunications.

Hamish was first licensed as VK3ZMV in Ballarat, Australia in 1960. Two meters was his main band of interest, followed later by 70 cm. His equipment was all home-built, mainly using surplus World War II parts. When reasonable solid-state devices appeared, he turned his interests to semi-portable operation with home-built rigs, and also published some articles in local newsletters covering TTL logic based frequency synthesizers and 2 meter solid state amplifiers. Hamish was absent from Amateur Radio for a while afterward until the early 2000s, when he took the Finnish Radio Amateurs examination and was licensed as OH2GAQ. He is now gradually putting together equipment for the various microwave bands including, particularly in a portable form, for the SHF bands.

Hamish is married and has five children. His wife is a building engineer. Much of his spare time is taken up with building projects for the family, including house renovations for the various grown-up children. During the summer he enjoys boating in the local archipelago in southern Finland, not to mention installing more electronic gadgets in the boat. Of course the next summer house project is always around the corner!

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Motion Based Electrical Power Control

Micro-electrical-mechanical systems (MEMS) ICs provide some interesting motion-based electrical power control possibilities. What other applications can you find for these devices?

MEMS Are Fun

MEMS (micro-electrical-mechanical systems) inertial sensors, rarely encountered by the general public a decade ago, have become commonplace recently. You may, in fact, use these sensors without knowing it. Some examples include the Wiimote controller on a Nintendo Wii, the chip inside your "smart phone" that knows which way is up and rotates the display, or the airbag deployment systems in your car.

MEMS sensors are becoming ubiquitous. We have only seen the tip of the iceberg in terms of the possibilities of enhancing things using MEMS accelerometers. This article shows how MEMS sensors can be used in Amateur Radio and can enhance mobile equipment or "Go Boxes."

Some of the first MEMS accelerometer-based applications I built demonstrated the digital features of the ADXL345 three axis digital accelerometer. I used its activity and inactivity detection and double-tap detection features to turn an LED on and off. I envisioned that the double-tap-detection device, which turned the light on or off when you tapped or hit it twice rapidly, could replace a power switch in corrosive or hazardous environments where an open switch would not last well. One example would be in a waterproof flashlight. I built another example, using activity and inactivity detection, where the light turned on when you picked it up and turned off shortly after it stopped moving. This might be a good idea for an emergency flashlight that turns on when you pick it up.

[1]Notes appear on page 25.

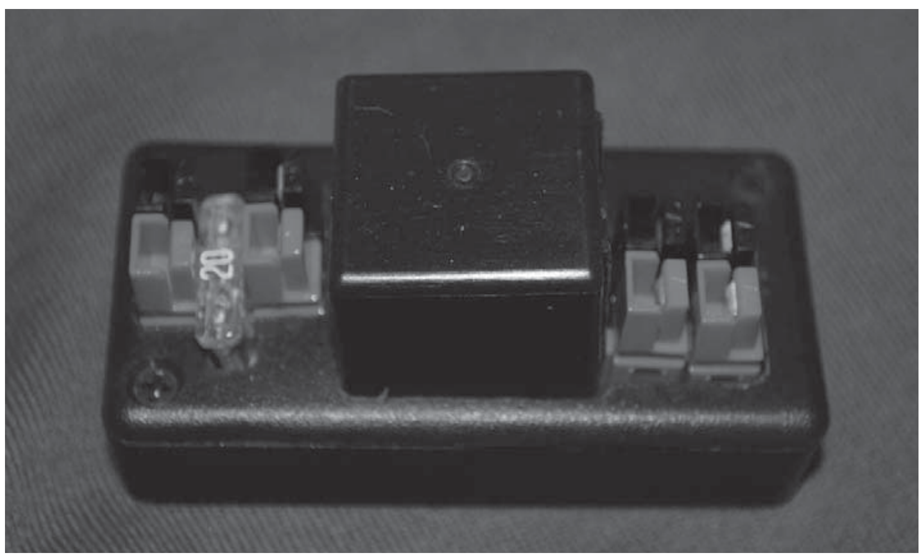

Figure 1 — Here is a photo of one complete motion switch that I built.

Motivation for a Radio Power Switch

I thought that the next step would be to build a little power switch in my car to turn the radios on and off automatically when the car starts to move. Figure 1 shows the complete switch. Although automatic switches that turn the radio on when the car is running and off when the engine is stopped already exist, these normally work by monitoring the car battery voltage. With hybrid-powered cars and trucks, this voltage sensing technique will not be effective. A motion activated switch is useful in all types of vehicles, particularly for portable and battery-operated equipment.

Some Potential Uses:

- Turn on two-way radios automatically.
- Turn on GPS navigation systems automatically.
- Turn on GPS (APRS) trackers automatically.
- Turn on equipment in a “Go Box” automatically.
- Turn on tracking equipment in other emergency equipment, such as medical supplies or AEDs.

The Need for Automatic Power Switches in a “Go Box”

A “Go Box” refers to a communications device intended for rapid deployment when portable or emergency communications is required. The box typically consists of communication equipment and batteries. The equipment may be left unused on a shelf for several months, but needs to be ready for use any time it is picked up off the shelf.

Go boxes typically have a master switch that should be in the **OFF** position while the equipment is not in use, but as always, the switch poses some challenges. A “master switch” inside the box may be left in the **ON** position, draining the batteries and rendering the equipment useless. It may be forgotten in an emergency, causing frustration for the user who doesn’t know how to turn the box on. A “master switch” on the outside of the box would have all the problems of an internal switch. In addition, it is more prone to breaking and may also accidentally switch when moved or when other equipment is moved next to it.

A motion-activated switch addresses all of the issues of a manual switch. The power turns off automatically when the Go Box is not in use, and will turn on automatically when it is moved for deployment. In addition, if the “Go Box” has a GPS tracker, it will be powered automatically each time the Go Box is moved and will be easier to track.

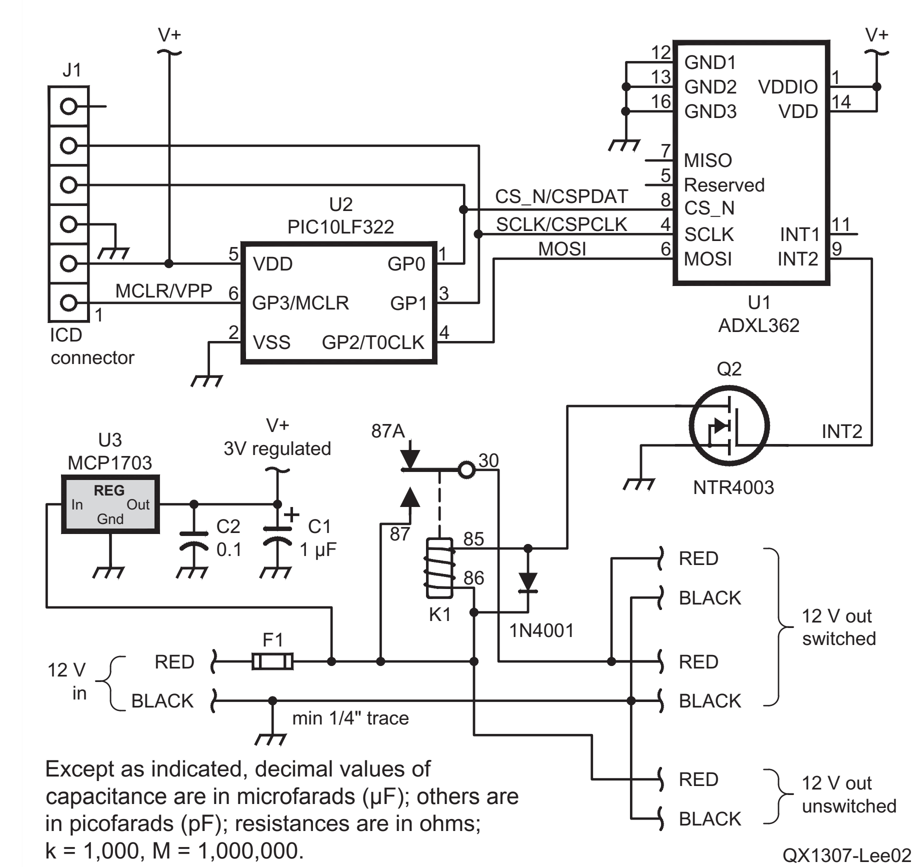

Figure 2 — This schematic diagram shows the circuit for my motion switch.

Table 1
SPI Bus Master/Slave Signals

Signal	*Direction*	*Description*
CSN	M->S	Active low chip select - Sometimes called SS (Slave Select). When this line is low, the sensor knows to listen.
CLK	M->S	Clock, anywhere from a few KHz to 10’s of MHZ, around 1 MHz used in this article.
MOSI	M->S	Master Out – Slave In – This is the data sent from the microprocessor to the sensor.
MISO	S->M	Slave Out – Master In – This is the data sent from the sensor to the microprocessor. In this example the data is *not* read, so this signal is *not* used.

Hardware

The ADXL362 micropower digital accelerometer has some improvements over the ADXL345, making it a better sensor for a motion power switch. It is lower power and adds improvements to the activity and inactivity detection functionality. I have used the ADXL362 to implement the motion switch discussed here.

Figure 2 shows a schematic diagram for the automated power switch. As you can see, it has an unswitched power connection that can be used for a charger or always-on device. It also includes two switched power connections that can be used to power a radio (or two) and a GPS.

The basic circuit consists of a PIC10LF322 microprocessor to initialize the accelerometer, the ADXL362 accelerometer, an FET and an automotive relay that switches the load.

Software

The program on the microprocessor is short, as shown in Figure 3, consisting of only a few SPI transfers.[1] The processor sends a few SPI commands to configure the ADXL362, and then sleeps. The magic is all in the ADXL362 and properly configuring it. The accelerometer is configured to have the AWAKE status come out the INT2 pin, which directly controls the FET and switches the relay.

The accelerometer is configured by the software shown in Figure 3 to detect the type of motion of interest to turn on the radios. The ADXL362 is configured to ignore the constant 1 g field of the earth. The configuration is set to look for about 105 mG change in acceleration for three samples to detect the start of motion. Once motion is detected it continues to stay on for 15 minutes past the last change in acceleration over 105 mG. This threshold of about 0.1 G is quite easy to achieve when moving. Since the thresholds and timing are all software controlled, it can be customized for other applications. The settings in this program are recommended

```
/*--------------------------------------------------
    File Name          :
    Author             : MPD Application Team
    Version            : V0.0.1
    Date               : 11/06/2008
    Description        :
    File ID            : $Id: xl362_tilt_main.c,v 1.1 2012/05/02 19:43:54 jlee11 Exp $

  -------------------------------------------------------*/
  //  include <pic.h>
  //  include <datalib.h>
  #include <htc.h>
  #include "XL362.h"
  #include "xl362_io.h"
  __CONFIG(FOSC_INTOSC &  BOREN_OFF & PWRTE_ON & CP_OFF & MCLRE_ON & LVP_ON & LPBOR_OFF & WDTE_OFF );

  void i2cinit(void);
  unsigned char buffer[8]; /* a buffer to use with the read and write functions */
  void main(){
    OPTION_REG = 0x87 /*disable wakeup and pull-ups max pre-scale*/;
    OSCCON = 0x30; // 1 mhz

   /* power up timer */
   buffer[3] = 255;
   while(buffer[3]--);
   /* initialize SPI with two device ID read */
   xl362Read(4,XL362_DEVID_AD,buffer);
   xl362Read(4,XL362_DEVID_AD,buffer);

   /* soft reset for safety */
    buffer[0] = XL362_SOFT_RESET_KEY;
    xl362Write(1,XL362_SOFT_RESET,buffer);

   /* wait for soft reset to pass */
   buffer[3] = 255;
   while(buffer[3]--);

    /* set up a buffer with all the initialization for activity and inactivity */
    buffer[0] = 105;  /* XL362_THRESH_ACTL about 15 degrees*/
    buffer[1] = 0  ;  /* THRESH_ACTH */
    buffer[2] = 3  ;  /* TIME_ACT */
    buffer[3] = 105;  /* THRESH_INACTL*/

    buffer[4] = 0  ;  /* THRESH_INACTH */
    buffer[5] = 0xf2 ;  /* TIME_INACTL 15 minutes at 12.5 hz*/
    buffer[6] = 0x2b  ;  /* TIME_INACTH */
    buffer[7] =     /* ACT_INACT_CTL */
             XL362_ACT_ENABLE | XL362_ACT_AC | XL362_INACT_ENABLE
      | XL362_INACT_AC | XL362_ACT_INACT_LINK | XL362_ACT_INACT_LOOP;
    xl362Write(8,XL362_THRESH_ACTL,buffer);

    /* set up a buffer with all the initiation for int maps filter and power*/
    buffer[0] = 0 ; /* INTMAP1 */
    buffer[1] = XL362_INT_AWAKE ; /* INTMAP2 */
    buffer[2] =     /* FILTER_CTL */
      XL362_RATE_12_5 |  XL362_RANGE_2G;
    buffer[3] =     /* POWER_CTL */
      XL362_MEASURE_3D |  XL362_LOW_POWER;
    xl362Write(4,XL362_INTMAP1,buffer);

    /* No interrupts INTCON = 0x90; */
    while(1){ /* we only resume here after a wakeup interrupt */
      asm("sleep"); /* go into low power mode */
    } /* while */
  } /* main */
```

Figure 3 — This listing gives the main *C* code for the motion switch.

since they have worked fine for the author.

Note that the xl362Write() function and the header file are omitted for clarity. The full software is available on the Analog Devices website, in the Engineer Zone.[2]

SPI Basics

The ADXL362 uses an SPI port to communicate with a microprocessor. The SPI bus is a simple 4 wire bus used to talk with peripheral chips. It has been used inside electronic devices for more than 20 years. On an SPI bus there are two sides of the bus; a master — typically a microprocessor — and a slave, in this case the MEMS sensor. Aside from power and ground the signals are listed in Table 1.

On an SPI transfer, data can be sent in both directions at the same time due to the MISO and MOSI lines being dedicated connections. The meaning of bytes sent on the SPI bus is up to individual component designers. For the ADXL362, it uses a packet-like format, where it receives a read or write command byte, a register address byte and then either sends or receives bytes of data starting from the specified address.

For the ADXL362, the xl362Write() routine just wraps this transaction into an easy to use format. Those transactions look like:

1) <CSN Down>
2) <send 0x0A Write Register command>
3) <send register address>
4) <Send data byte(s)>
5) <CSN Up>

Since the microprocessor used does not have a hardware SPI-port the included library code xl362_io_pic.c uses software to wiggle the clock and data signals directly. At the heart of each of the transfers is the xfer_byte() routine, which wiggles the clock and simultaneously sends data on MOSI and reads the data on MISO. All the larger SPI transactions just use xfer_byte() and know what to tell it to send or listen to its result. More details on the SPI protocol and registers can be found on the ADXL362 data sheet.

The attached code was written by me and provided by Analog Devices and published for public use in their electronic customer support system.

Putting It All Together

The *C* code was compiled on Microchip's MPLAB IDE™. Although I used MPLAB8, I have also converted the code to compile under MPLABX™. Once the code was compiled, the PICKIT3™ programmer was connected to the board and configured to power the board from the PICKIT3. (MPLAB IDE™, MPLABX™, and PICKIT3™ are all Microchip trade marks.)

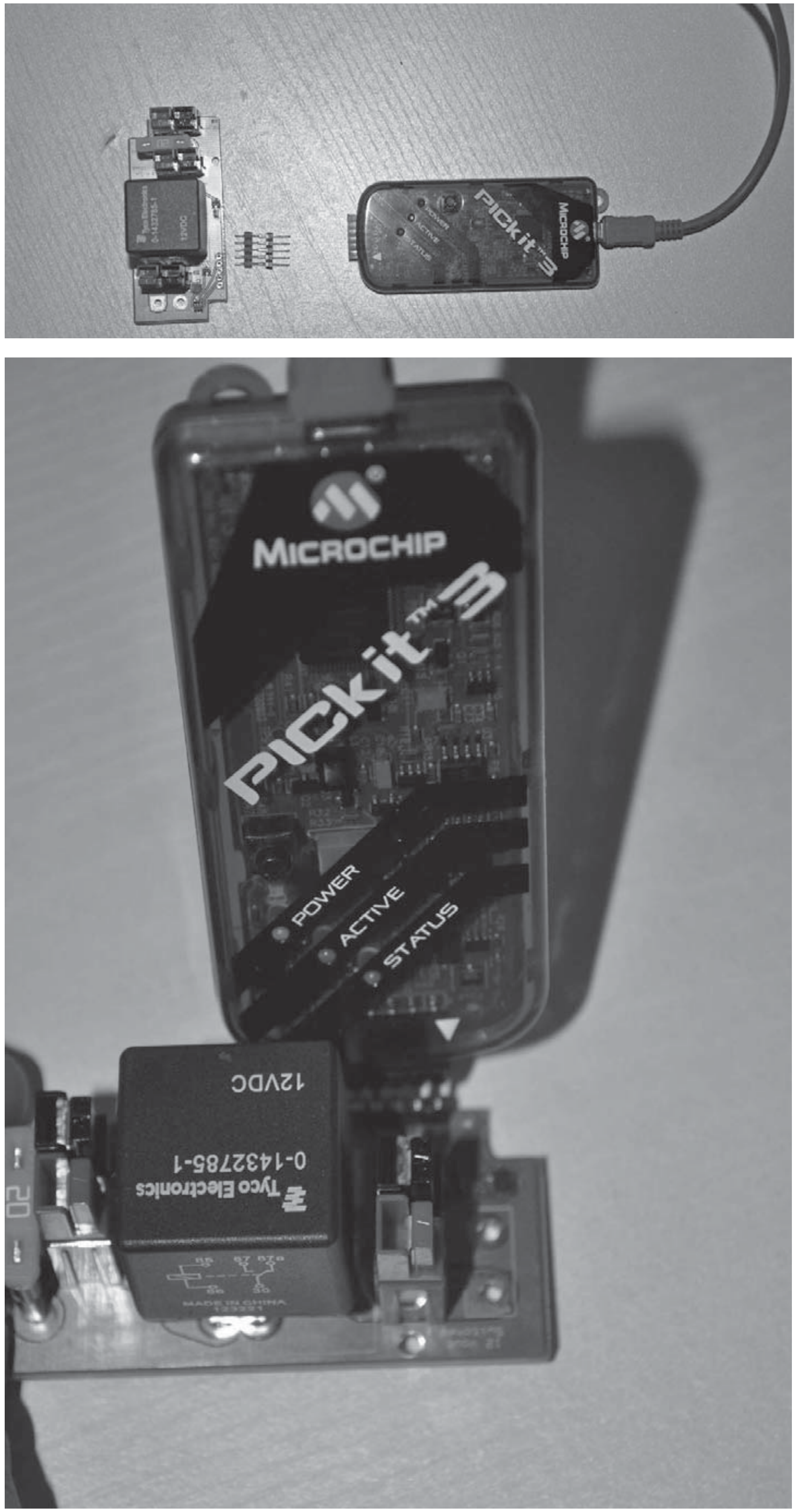

Figure 4 — Part A shows the assembled switch, the header pins and the PICKIT ready to plug in and program the PIC10LF322 microprocessor. Part B shows how the PICKIT with header pins plugged into the switch circuit board while the PIC microprocessor was programmed.

We built an adapter to convert the female pins on the PICKIT3 to male pins and then just placed it into the holes on the board and the connections were good enough without a socket or solder to program the processor. Figure 4 shows the pieces and the assembly to program the board.

MEMS Activity Sensing 101

MEMS accelerometers such as the ADXL345 and ADXL362 can measure acceleration (or gravity) on all three axes. Human movements generally have bandwidths ranging from under 0.5 Hz to about 3 Hz. For example, an Olympic runner may move their legs at a peak of about 6 Hz.

If we look only at acceleration due to gravity we can determine orientation. For a motion switch, we are not interested in orientation. We want to look at changes in acceleration to know something has been moved. We also want to suppress short duration movement such as bumping the "Go Box" or closing a car door. The ADXL362 allows the activity and inactivity detection to be based on changes in acceleration only, eliminating the effect of gravity, and also allows using a 6 Hz low pass filter.

Power Consumption

One commercially available Anderson Powerpole based voltage sensing switch consumes nearly 2 mA, significantly more than the current used by this automated switch. Obviously, completely passive panels with only fuses, switches and connectors should not consume any power. Those with LED status indicators, voltage monitors and even just protection diodes will have some current consumption, however, even if it is just the leakage current of the protection diode.

With any battery operated device — even those in a vehicle — we should consider the leakage or standby current. The standby current is the current that drains the battery while the device is not in use. You don't want to come and find a dead battery when you need to use it. In Amateur Radio, we typically measure current consumption in amps (A) or perhaps tens of milliamps (mA). This design drops current consumption to a whole new level — down into the microamp (μA) range!

Other than the relay, all of the circuitry runs on 3 V, but the application is intended to run on a 12 V battery. The first step was to select a voltage regulator. Many voltage regulators draw more than 75 μA with no load. Since the heart of this system is the ADXL362, which draws around 2 μA, using a regulator that draws significantly more power than the accelerometer was out of the question. The PIC processor used in this circuit takes about 1 μA in sleep mode (recall that it is in sleep mode for all but a few milliseconds when power is first applied to the system), the ADXL362 about 2 μA, and the selected regulator about 2 μA. This adds up to less than 10 μA, about 1/1000 of what an LED takes to light. Once power is on and the radios are on, the power consumed by the relay is assumed to be inconsequential.

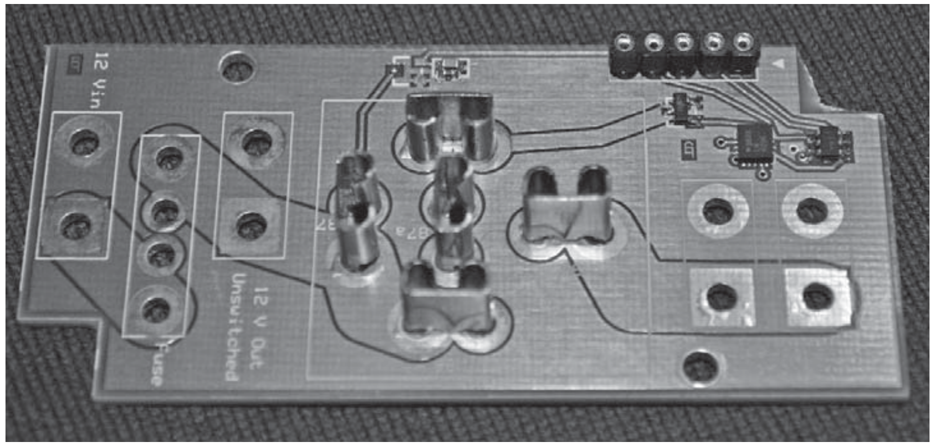

Figure 5 — This photo shows the switch circuit board with the relay mounting pins and the FET, ADXL362 accelerometer, and the PIC10LF322 microprocessor mounted to the board. The large holes at either end are for the Anderson Powerpole connectors and the fuse block.

Table 2
Bill of Materials

Part Number	*Vendor*	*Description*	*Quantity*
HM377-nd	Digikey	Hammond box	1
MCP1703T-3002E/CBCT-ND	Digikey	3-V regulator	1
NTR4003NT3GOSCT-ND	Digikey	FET	1
3525K-ND	Digikey	faston clip (for fuse)	2
3528K-ND	Digikey	faston clip (for relay)	5
PB680-ND	Digikey	Relay	1
F4201-ND	Digikey	20 Amp Fuse	1
PP10-30	Powerwerx	Anderson Powerpole connectors	4 pair
ADXL362 BCCZ-R2CT-ND	Digikey	Low power accelerometer	1
PIC10LF322T-I/O TC-T-ND	Digikey	PIC Processor	1
		Printed Circuit Board	1
		2-56×¼ screws	2

Building the Prototype

I fabricated a circuit board for the prototypes. Since the board needs to be able to carry high current to the loads, I used wide traces on both sides of the boards. I also selected appropriate clips to make sockets for the fuse and relay. I chose a Hammond 1551K box in which to mount the assembly. The fuse and relay are easily attained at an auto parts store. This makes replacement easier if anything ever needs servicing. (See the bill of materials.)

The surface mount parts were soldered directly to the board. The clips for the relay and fuse were placed on the relay and fuse before soldering to the board to help alignment. To solder the Powerpole connectors to the board, I crimped a short length of wire into the connectors and then soldered them to the board.

Figure 5 shows a partially assembled board. On the left you can see the holes for the 12 V input power, the fuse, and the unswitched power output. In the center are the clips for the relay. Above the relay are the pads for the 3 V regulator and bypass capacitors. On the top right is the connector used for programming the PIC. Below the programming connector, from left to right, are the FET used to switch the relay, the ADXL362 accelerometer, and the PIC10LF322 microprocessor. Below these active components are the holes for the switched power outlets.

Testing and Using the Switch

The switch was installed in my car (see Figure 6) and used to switch on a radio and a GPS. I noticed that the radio turned on just after the car started moving. I decided this was actually a nice bonus that the turn on

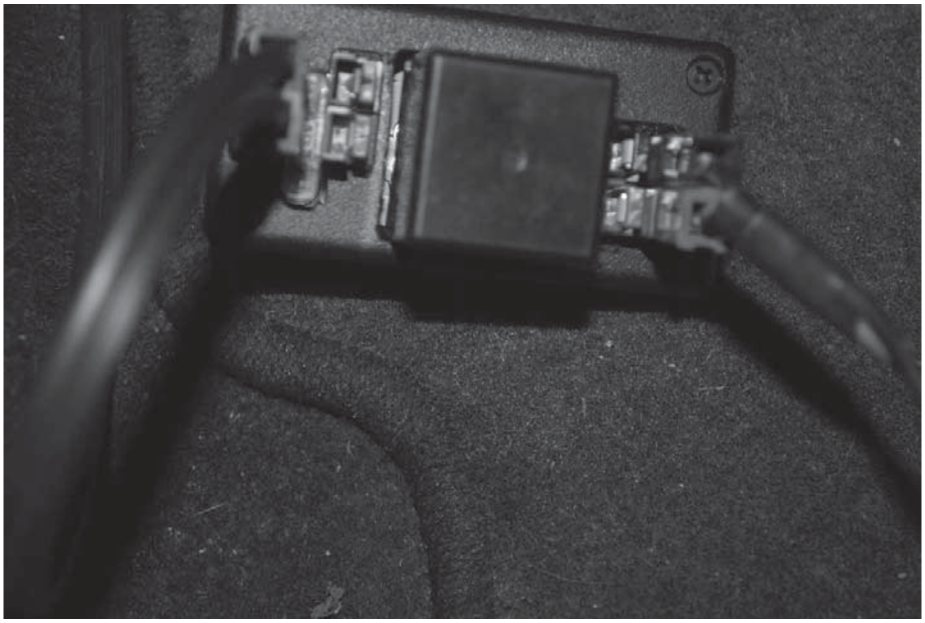

Figure 6 — Here is the automatic switch Installed in my car.

transient for the radio and GPS happened after I started the car. If I drove very gently in the parking lot, the switch stayed off until I hit the gas or brake. The switch never turned off while I was driving and I found the results to be just as I wanted. Resulting GPS tracks were observed on **findu.com**.

When I say that I "installed" the switch, I am using that term loosely — it is just sitting on the floor. Because the switch automatically detects motion and is not concerned with its base orientation, it can be installed anywhere or just left sitting on the floor or under a seat. With it sitting on the floor you can "jiggle" it at any time to test it and ensure that it turns on the power.

In Conclusion

It works; it's a fun project; go build one. In a "Go Box" on a 7 Ah battery (one of the most common gel cells used by Amateur Radio operators), the motion switch would theoretically take about 80 years to discharge the battery. In reality, the self-discharge rate of the battery is much greater, so the power draw of the motion switch is negligible. This simple circuit can protect your battery by keeping everything off until you need it. I encourage experimenters and builders to reproduce this project and to find other applications for MEMs devices. Please contact me for commercial production details.

Thank you to Joseph Lath for his help in laying out and building the prototype circuit boards.

James Lee, N1DDK, originally licensed in 1982, holds an Amateur Extra class license. He is the author of Verilog Quickstart*, now in its third edition. James is an ASIC and FPGA Design and Verification engineer. He has been designing sensor and microprocessor based systems for over 25 years. He has designed semiconductors for many companies and was the digital designer of the ADXL345 and ADXL362. James is designing chips for Qualcomm.*

Notes

[1]The program code and other files for this article are available for download from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **7x13_Lee.zip**.

[2]You can find the complete software code and more information about using the ADXL362 to eliminate a power switch and control system power on the Analog Devices website, in the EngineerZone section at **ez.analog.com/message/40001**.

What are MEMS?

MEMS is short for Micro-Electrical-Mechanical-Systems.

The largest segment of MEMS devices are micro-fluidic parts (miniature pipes, pumps and valves) used in ink-jet print heads. Micro-fluidic parts are also used in so-called "lab-on-a-chip" used in advanced medical devices.

Another well known MEMS segment is micro-mirrors. Micro-mirrors are used in digital light processing (DLP) chips used in some projectors used in home, office and even movie theaters.

Some micro machines are gears and motors. Others are microscopic relays called MEMS switches often used as antenna switches and band switching in small RF circuits in cell phones. MEMS resonators are replacing crystal resonators in some oscillator systems.

In this article MEMS inertial sensors such as accelerometers and gyros are discussed. These are systems of microscopic weights and springs along with circuitry to measure their displacement. MEMS microphones are similar to inertial sensors but instead of a mass being measured there is a diaphragm. MEMS microphones are used in many cell phones and tablets due to their small size.

MEMS components use many different materials and manufacturing methods. A MEMS device can be as simple as a conductive ink on a piece of paper so the resistance changes in response to stress (bending and flexing). There are also piezo MEMS devices, which can generate a small voltage or change in resistance when used as a sensor, or change shape when a small voltage is applied. Some MEMS devices are capacitive and generate a change in capacitance when a mechanical stimulus is applied. The accelerometer used in this design is a capacitive device. Some MEMS devices are electrostatic and move when a voltage is applied. MEMS Switches, DLP, and micro pumps and valves are generally electrostatic.

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Using Time Domain Reflectometry for Transmission Line Impedance Measurement

Ray revisits a technique for transmission line measurements using modern techniques.

A Little Background

This is not the first article about time domain reflectometers (TDR) in the Amateur Radio press. The earliest one I found was in 1973 and then again in 1989 both in *QST*.[1, 2] There were also four articles in *Ham Radio* during the 70s and 80s.[3, 4, 5, 6] In the Mar/Apr 2006 issue of *QEX* there is an excellent article about a pulse generator and its application as a TDR.[7]

I am doing some research on transmission line transformer construction, so I needed a way to verify the actual impedance of custom transmission lines. I have an MFJ-259B Antenna Analyzer, but it is a frequency domain device and assumes a 50 Ω system. You could use such an instrument by tuning for zero impedance with a ¼ λ open circuit feed line, an infinite impedance at ½ λ open circuit feed line, and a transformed impedance with a known load. Such measurements require knowing the characteristics of the line under test, such as velocity factor and impedance, however. Since line impedance is the primary characteristic I need to measure, I require a different tool.

Vector network analyzers frequently have tools for doing such measurements, but my current employer has no such device and neither do any friends for whom it would be convenient. I remembered using the Tektronix TDR in my sophomore EE lab back in the 70s. Such a tool would be perfect for measuring impedances. It turns out that a TDR is just a high speed oscilloscope connected with a fast rise time pulse generator, and I have access to both. Figure 1 shows the connections to create a TDR.

[1]Notes appear on page 30.

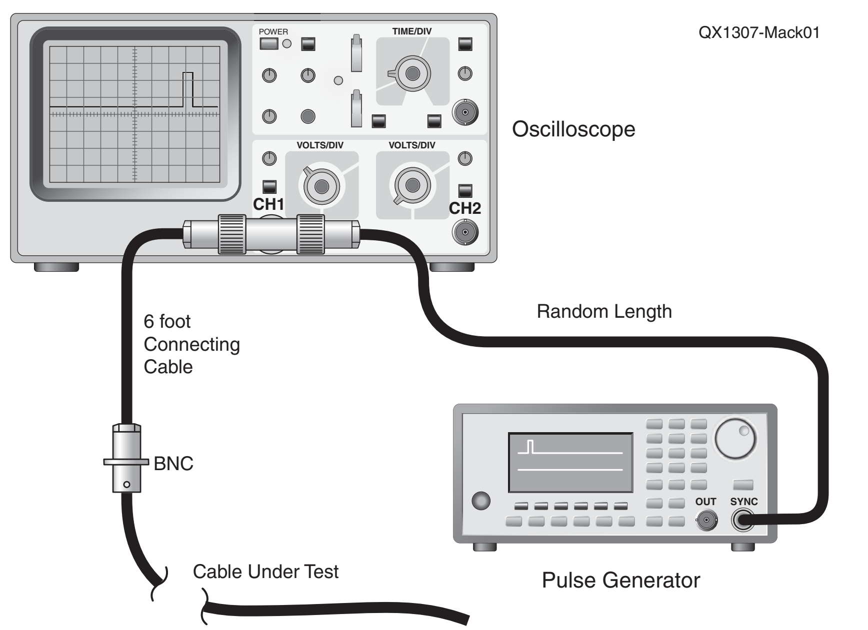

Figure 1 – The connections to turn an oscilloscope and a pulse generator into a time domain reflectometer.

TDR Fundamentals

A TDR is a traveling wave measurement device. It uses the property that a traveling wave will enter a transmission line with the transmission line appearing as a resistor whose value is Z_0 of the transmission line. Figure 2 shows the analogy when the transmission line is infinite length and it starts with zero voltage across terminals A and B. Turning on the switch in either circuit will result in 2.5 V across terminals A and B and

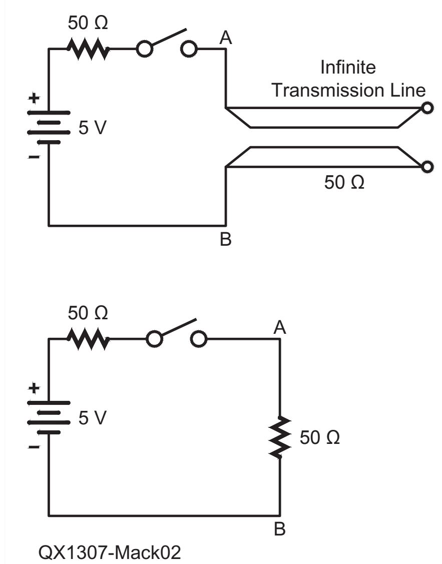

Figure 2 – A comparison of a resistor network and a source driving an infinite length transmission line.

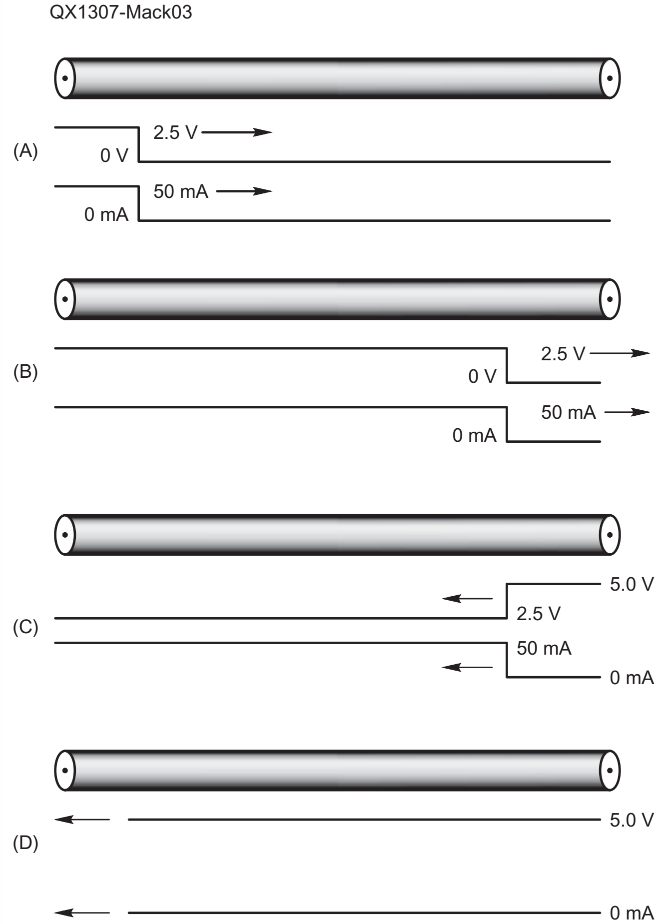

Figure 3 – Part A shows the wave shortly after being launched toward the far end. Part B shows the voltage and current along the transmission line as the wave approaches the end of the cable. Part C shows the returning wave just after the incident wave reached the end of the line. Part D shows the result after the reflected wave reaches the source.

50 mA of current flowing out of terminal A. In the case of the transmission line, a wave front of a 2.5 V step begins propagating down the line at the wave velocity of the line. This value is the speed of light times the velocity factor. This process continues in our infinite line as the 50 mA of current adds charge along the length of the transmission line to create a 2.5 V field as time increases.

We don't have infinite length transmission lines in the real world, so what happens when the wave hits the end of the line? Figure 3 contains our example. Let's assume that the line is an open circuit. As soon as our traveling wave step function hits the open circuit the 50 mA of current has nowhere to go. In order for the 50 mA to go to 0 mA, the line has to create a negative 50 mA of current. This negative 50 mA of current is applied across the same 50 Ω impedance of the line, but the direction is opposite. The result is that the –50 mA times 50 Ω creates a 2.5 V step that is in phase with the arriving 2.5 V. The voltages add to give us 5.0 V and the currents subtract to give us 0 mA. This 2.5 V / –50 mA wave propagates back towards the source, leaving in its wake a 5.0 V charge on the line and net zero current.

If you look at the input side of the line at the time the reverse wave starts to travel, the voltage source is still supplying 50 mA of current that is flowing into the line and 2.5 V is still applied across the line. The –50 mA in essence gobbles up the 50 mA that is still being pumped into the line as it travels back towards the source. Once the returning wave gets to terminals A and B, we have a net zero current and 5.0 V, which is exactly what we would have if there had been an open circuit at terminals A and B instead of having the open circuit at the far end of the line.

A similar situation occurs if the far end of the line is shorted. See Figure 4. When the 2.5 V / 50 mA wave reaches the short circuit, the voltage must be zero. That requires that a –2.5 V value be added to the incoming voltage. This –2.5 V across the 50 Ω line impedance generates +50 mA, which adds to the incoming 50 mA resulting in 100 mA of current. When this wave gets back to the source, the voltage is zero and the current is 100 mA just as it would have been if the short were at the terminals A and B.

Similar situations occur for resistances other than a short or open that are attached at the end of the transmission line. Any resistance that is higher than Z_0 will cause a higher voltage to be reflected and any lower resistance will cause a lower voltage to be reflected. We can predict the actual voltage on the line for the return step by using the voltage divider equation:

$$V = V_{in}\left(\frac{Z_l}{Z_0 + Z_l}\right) \qquad \text{[Eq 1]}$$

It is important that the source impedance is equal (or at least close) to the line impedance or we will create an infinite sequence of waves that bounce back and forth from one end of the line to the other pretty much forever, and result in a situation similar to standing waves with sine wave excitation. In fact, TDR operation is a standing wave at zero hertz once the initial reflection occurs.

Classic TDR Use

The classic TDR application is to look for issues in long transmission lines especially where it would be difficult to physically inspect the line. Figure 5 is the TDR display of my 2 m feed line connected to my 2 m Yagi. My guess was that it is approximately 50 feet of LMR400 followed by about 30 feet of RG8X and then about 16 feet of LMR400

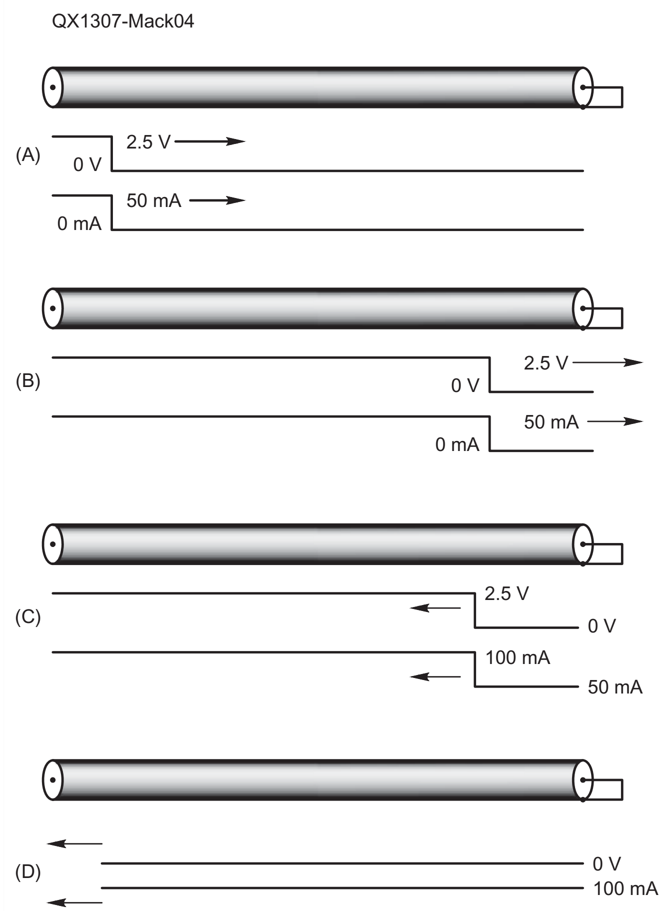

Figure 4 – The diagrams of the incident and reflected waves when the transmission line is shorted instead of open. Notice that the voltage waveform is inverted from that in Figure 3.

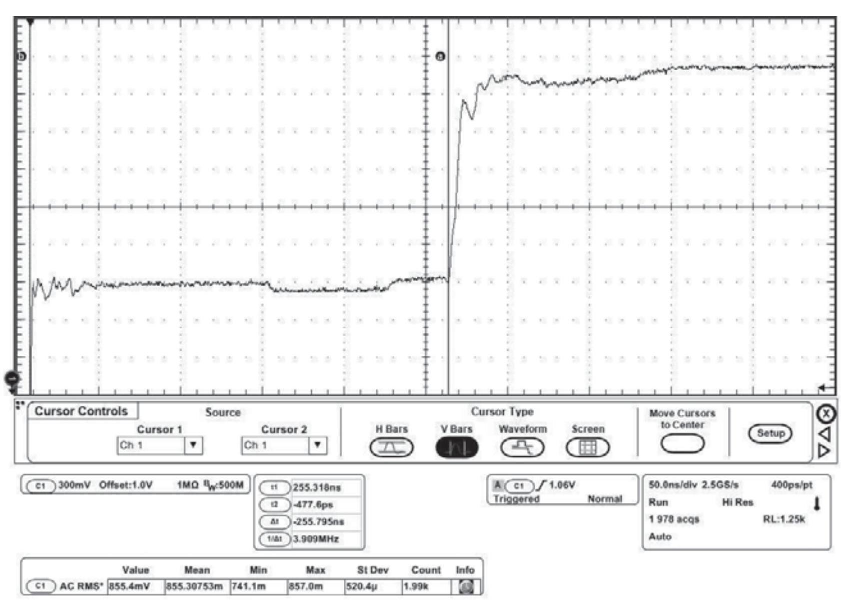

Figure 5 – The TDR measurement of my 2 m feed line with a mix of LMR400 and RG8X coaxial cables.

ultraflex. The velocity factor of LMR400 is specified at 0.85 and RG8X is 0.82. Times Microwave makes our task even easier by supplying the inverse property of 1.2 ns/ft transit time. RG8X corresponds to 1.24 ns/ft. So this run of feed line should take about 160 ns for the two pieces of LMR400 and 74 ns for the RG8X or 232 ns for the round trip. It turns out that I was off by a few feet based on the TDR. You can clearly see the impedance bump of the RG8X at just about 120 ns and a second bump at 152 ns. So, my cable run is really 50 feet of LMR400, 30 feet of RG8X, and 20 more feet of LMR400 ultraflex. Figure 5 also shows that the RG8X is slightly lower impedance than the LMR400 and the LMR400 ultraflex is slightly higher impedance than the other two.

I took a measurement before beginning to write this article that looked just like Figure 5 but did not capture it. I got an object lesson in TDR use when I went back to capture the display. I got Figure 6 instead of Figure 5. It clearly shows a problem with my temporary connection at the top of the tower. I went up the tower and tightened the two PL259 connectors onto the barrel connector. The RG-8X connector had twisted loose in the wind over the course of the week. Figure 5 is the result after the repair at the top of the tower.

Oscilloscope Features

Figures 5 and 6 were taken using an expensive MSO5104 scope that has vertical input of 1 MΩ in parallel with 13 pF with 1 GHz bandwidth and 10 GS/s. The scope is being driven by a HP33120 Arbitrary Waveform generator at 1.2 kHz. The ringing that you see at the rising edge of the trace occurs from reflections on the short cable between the pulse generator and scope. The 13 pF capacitance of the scope input causes an impedance bump. It is very difficult to remove the ringing due to the scope capacitance. This scope is excellent for the measurements because you can subtract a voltage from the baseline and increase the volts per division to increase the resolution. Unless you are lucky enough to have an employer that allows use of such scopes, you are more likely to be able to use a scope such as the MSO2012 that produced the rest of the scope traces for this article. This scope has 100 MHz bandwidth with 1 GS/s. It will measure time down to 4 ns/division and only costs approximately $1290 for the TDS2012 version. I have also used a much older TDS310 scope with 50 MHz bandwidth, 200 MS/s scope that can resolve down to 10 ns/division (but it does not have a usable screen capture function). A storage or digital scope is not a requirement. An older analog scope that can resolve 4 ns or less is also a possibility, but you will need to run

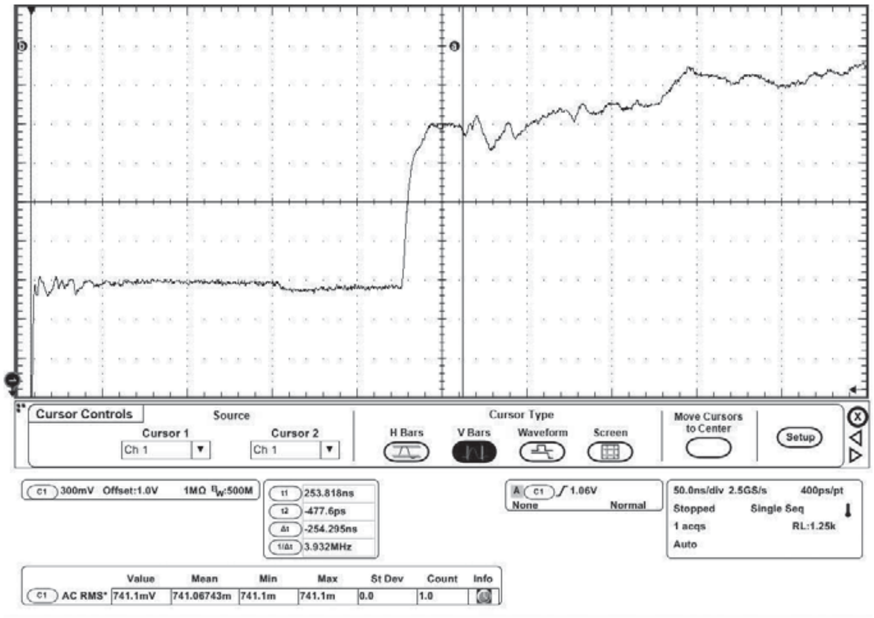

Figure 6 – The TDR measurement of my 2 m feed line when the PL259 at the far end of the RG8X was loose from vibration in the wind. This illustrates the classic use of TDR for finding where in a feed line there is a defect.

the pulse generator at a higher frequency to have enough trace brightness. Higher pulse rates will reduce the distance that you can measure, though.

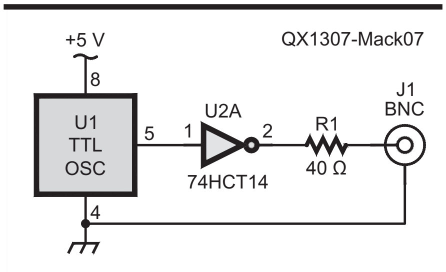

Figure 7 – The schematic of a fast rise time pulse generator for TDR use.

Characteristic Impedance Measurement

My problem is slightly different. I need to design optimal transmission lines for transmission line transformers with the lines being on the order of 18 to 36 inches long. For such short lines, the resolution needs to be the best possible both for the vertical and the horizontal systems. Figure 1 shows my equipment set up. The HP33120 has two pulse outputs. The main output has 50 Ω impedance but has a fairly long rise time specified at less than 20 ns. The better output is the sync output which has a TTL level (4.2 V open circuit), approximately 77 Ω impedance, and 10 ns rise time. These characteristics are close enough to use for the impedance measurement application. Tom King used a similar set up in his 1989 article. His system, likewise, does not have a 50 Ω source impedance. It is likely that the impedance is more on the order of 33 Ω.

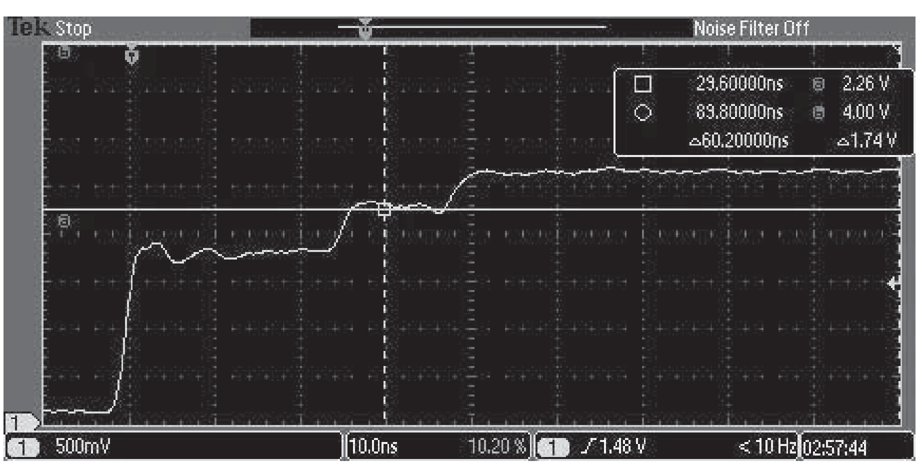

Figure 8 – TDR used to measure the impedance of 2 feet of CAT-6 twisted pair.

Table 1
Attenuation Versus Load Impedance for Attenuators Used as Precision Loads

Attenuation	*Load*
3 dB	151
4 dB	116
5 dB	96
6 dB	83
3 dB + 3 dB	75
10 dB	62.3
15 dB	53.5
30 dB	50
6 dB + 6dB	41.5
3 dB + 30 dB	37.6
30 dB + 30 dB	25
30 dB + 10 dB	27.7

Figure 7 shows the schematic of an alternate pulse circuit that will have a short rise time and close to 50 Ω impedance. The 74HCT14 is a Schmidt trigger gate that will convert just about any input signal into a fast rise time signal with less than 10 ns rise time. Any frequency 1 MHz or lower will work. A 1 MHz oscillator will give you the ability to measure at least 300 feet of cable with the 500 ns pulse time. You need to place the series resistor as close as possible to the 74HCT14 and also to the BNC connector. The resistor is labeled as 40 Ω, but that is an approximate value. The resistance will need to be adjusted so that the combination of the 74HCT14 output impedance plus the resistor is 50 Ω. First measure the open circuit voltage. It should be close to 5.0 V. You can place a 30 dB attenuator on the output as an accurate 50 Ω load. The output voltage will be one half the open circuit voltage when the source impedance is 50 Ω.

The final piece is a set of calibrated loads with minimal capacitance or inductance. The best tool for the purpose is a set of BNC attenuators used with one port open circuit. A BNC T connector allows combinations of loads in parallel, again with minimal inductance or capacitance. Table 1 shows attenuation with associated total termination impedance. Mini-Circuits sells their HAT series for $9.95 each, so a set of useful loads will set you back $50 (3 dB, 3 dB, 10 dB, 30 dB, 30 dB). The rise time and transit time are so short that even the minimal inductance of a rheostat or resistance substitution box will affect the measurements if you try to use them as a load. It helps to use a load as close as possible to the actual line impedance to reduce the errors due to reflections of an open circuit.

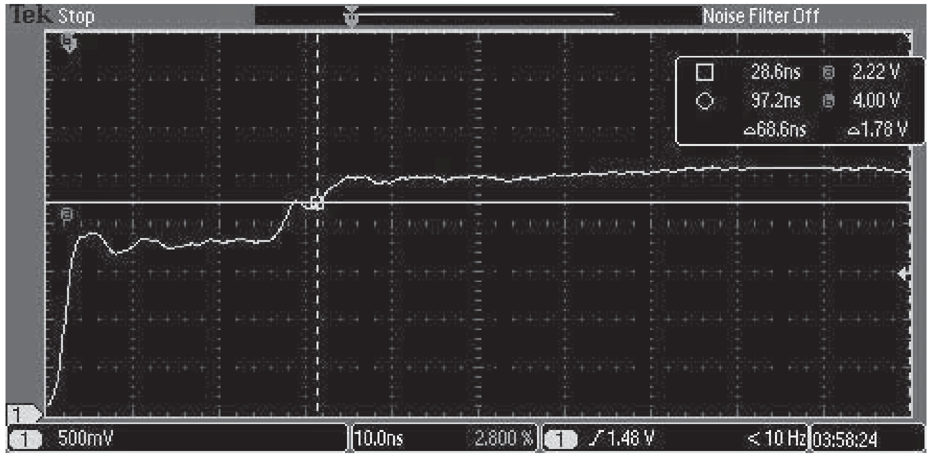

Figure 9 – TDR used to measure the impedance of 25 cm of two pieces of rigid 50 Ω Hardline in series, terminated in 150 Ω (30 dB attenuator). Note that the reflection is extremely short.

Figure 8 shows measurements for one of my test cables. The picture shows the results using a twisted pair from a 2 foot piece of CAT-6 cable. The specification for the impedance is 100 Ω, but the TDR measurement indicates it is really about 103 Ω. Figure 9 shows the result of two pieces of semi-rigid 50 Ω line in series. These results agree well with the expected value. The measurement shows 99 Ω.

Building a Precision TDR

Brian Straup, N5YC, and I discussed making a precision TDR. The simple TDR devices I presented above are limited by two factors: rise time of the pulse, and capacitance of the scope vertical amplifier. Brian's suggestion for short rise time is to use a microwave snap diode to generate an extremely short rise time pulse. Snap diodes are very good at this and are used for microwave comb generators. The second issue is the scope input capacitance. The solution is to use a very low capacitance FET as a buffer between the cable and the scope. This method is also used by scope manufacturers for the input of very high speed active probes. Such an amplifier can have less than 1 pF capacitance. These two enhancements will improve the response of a home built TDR.

Notes

[1]Warren Jochem, WB2IPF, "An Inexpensive Time-Domain Reflectometer," Mar 1973 *QST,* pp 19-21.

[2]Tom King, KD5HM, "A Practical Time-Domain Reflectometer," May 1989 *QST,* pp 22-24.

[3]Joe Carr, K4IPV, "Build Your Own Time-Domain Reflectometer (Practically Speaking)," Jun 1987 *Ham Radio* pp 69, 71-72.

[4]Bill Unger, VE3EFC, "A Time Domain Reflectometer," Nov 1983 *Ham Radio,* pp 49-51.

[5]Carl D. Gregory, K8CG, "Checking Transmission Lines with Time-Domain Reflectometry," Jul 1980 *Ham Radio,* pp 32-34.

[6]David M. Allen, WA0PIA, "A Practical Experimenter's Approach to Time-Domain Reflectometry," May 1971 *Ham Radio,* pp 22-27.

[7]Gary Steinbaugh, AF8L, "Pizzicato Pulse Generator," Mar/Apr 2006 *QEX*, pp 3-8.

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A Two Meter APRS Beacon Transmitter

A flexible transmitter that's ideal for experimenters.

This article describes a 2 W, 2 meter Automatic Packet Reporting System (APRS) beacon transmitter that reports GPS position at regular time intervals. Major parameters (call sign, APRS symbol, and so on) are set via RS232 and saved in nonvolatile EE memory. It requires an external GPS with RS232 output of NMEA GPRMC messages.

This design uses an Analog Devices ADF7012 transmitter chip, driving a Philips PD85004 RF FET, with a PIC 18F442 microcomputer for supervision, packet generation and control. The size is 1.2 × 2.2 inches and weighs just 0.24 ounces. The RF output at 12.6 V dc is 2 W, with a transmit dc load of approximately 400 mA (15 mA between transmissions). The PIC assembly code, CAD files and more are available on the *QEX* files website at **www.arrl.org/qexfiles**.[1] The schematic diagram is shown in Figure 1.

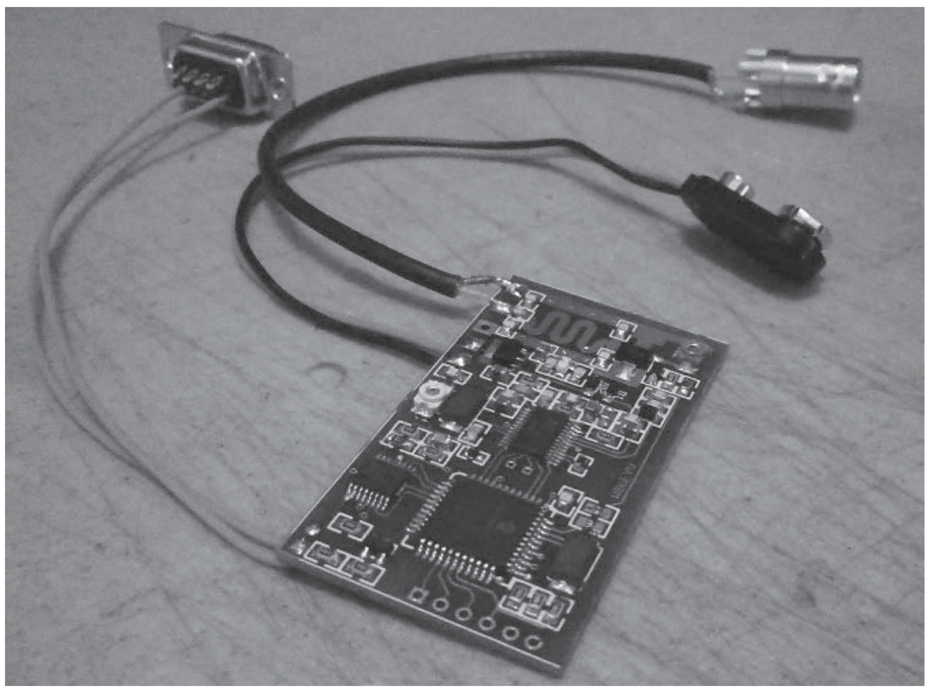

Photo A — Here is the completed beacon transmitter circuit board with programming, power and antenna cables connected.

Construction

Surface mount technology is employed throughout the design. The PIC chip is a TQFP package with a pin spacing of 0.030 inches, and the ADF7012 chip is an SSOP package with a pin spacing of 0.025 inches. Most other components are 0805 packages (0.080 × 0.050 inches), but the varicap diode is pretty tiny. The use of low profile components (including the two crystals) yields a transmitter with an overall profile height of only 0.13 inch.

The method I use to install ICs like these is best described as "glob and wick." Opposite diagonal (corner) pins of a chip are tacked down with solder to hold the chip in the proper position on the circuit board, followed by a close inspection to verify the pins are properly centered on the board footprint. The remaining two diagonal corners are then tacked down with solder to fully immobilize the chip.

A shameless glob of solder is then applied liberally to all the pins (shorting them all together) and allowed to cool. The excess solder is then wicked away using a clean section of solder wick laid across all the pins and heated so that *all* pins are simultaneously "wicked clean" by the capillary action of the wick. If everything is clean, the result is consistent and reliable and not too difficult. The gap between pins should be examined (closely) to verify no solder bridges exist. If bridges are detected, the pins can be "re-globbed" and wicked again.

For Experimenters

This beacon operates on 2 meters, but the basic design is very flexible and (in theory) can be adapted for operation on other bands from 100 to 1000 MHz. Provisions are made for experimenters; all unused pins on the PIC micro are terminated into uncommitted

[1]Notes appear on page 36.

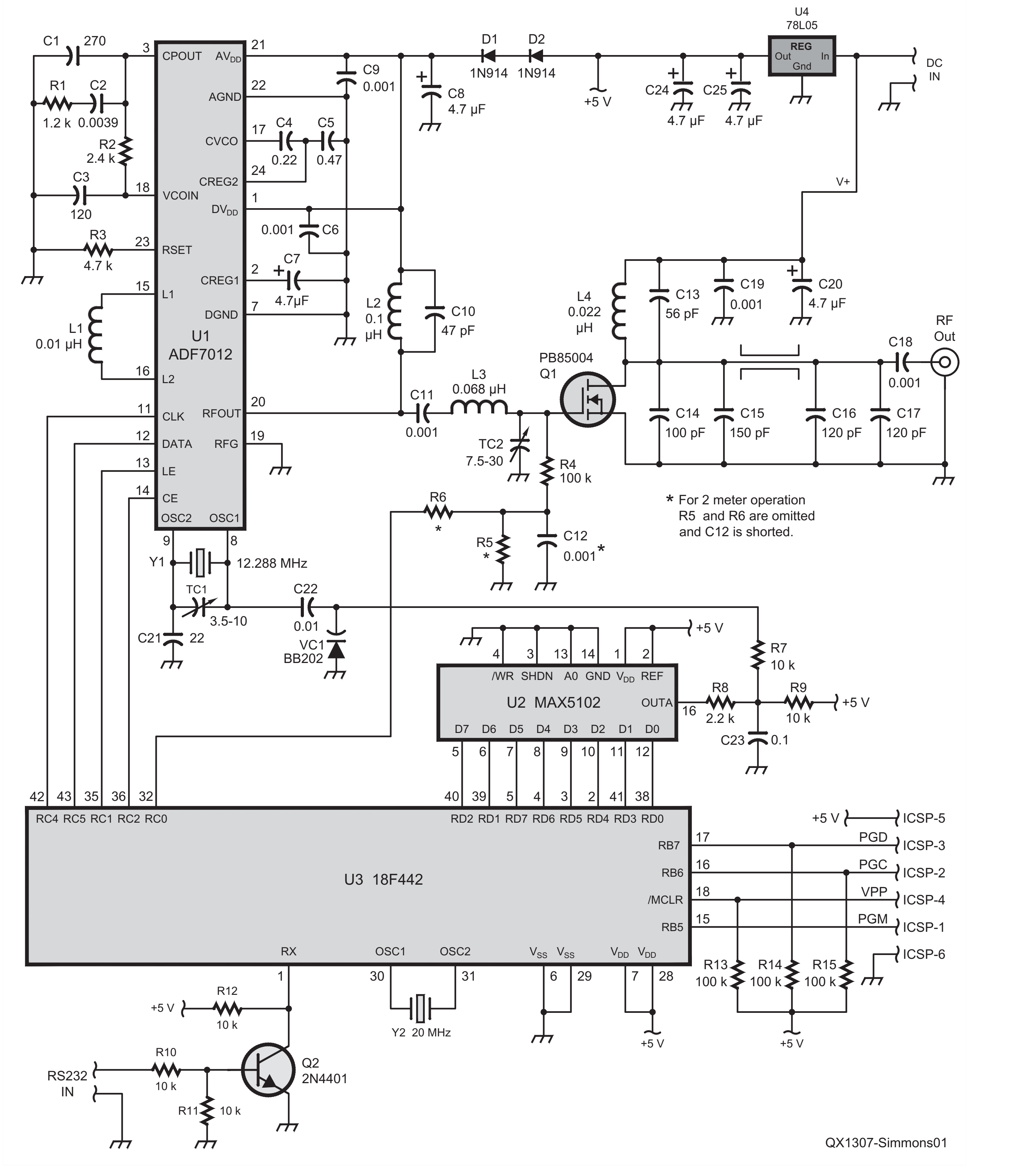

Figure 1 — This schematic diagram shows the beacon transmitter.

Parts List

Quantity	*Designation*	*Description*	*Part Number*	*Cost Each ($)*	*Extended Cost ($)*
1	n/a	Circuit Board (ExpressPCB)	Microbeacon.Board	11.00	11.00
1	U1	ADF7012	ADF7012BRUZ-ND	4.03	4.03
		ALT (Avnet)	ADF7012BRUZ	3.33	
1	U2	MAX5102	MAX5102BEUE+-ND	3.68	3.68
1	U3	18F442	PIC18F442-I/PT-ND	8.74	8.74
1	U4	78L05	296-13531-1-ND	0.38	0.38
1	Q1	PD85004	497-8291-ND	3.51	3.51
1	Q2	MMBT4401	MMBT4401FSCT-ND	0.21	0.21
2	D1, D2	1N4148	1N4148WSFSCT-ND	0.15	0.30
1	VC1	BB202	568-1953-1-ND	0.50	0.50
1	Y1	12.288 CS10	535-9837-1-ND	1.31	1.31
		ALT	XC1660CT-ND	1.69	
1	Y2	20.000 CS10	535-9843-1-ND	1.31	1.31
		ALT	XC1667CT-ND	1.69	
1	R1	1.2 kΩ 0805	P1.20KCCT-ND	0.10	0.10
1	R8	2.2 kΩ 0805	P2.20KCCT-ND	0.10	0.10
1	R2	2.4 kΩ 0805	P2.40KCCT-ND	0.10	0.10
1	R3	4.7 kΩ 0805	P4.70KCCT-ND	0.10	0.10
5	R7, R9, R10, R11, R12	10 kΩ 0805	P10.0KCCT-ND	0.10	0.50
4	R4, R13, R14, R15	100 kΩ 0805	P100KCCT-ND	0.10	0.40
1	C21	22 pF 0805	311-1103-1-ND	0.10	0.10
1	C10	47 pF 0805	311-1107-1-ND	0.10	0.10
1	C13	56 pF 0805	311-1108-1-ND	0.10	0.10
1	C14	100 pF 0805	311-1111-1-ND	0.10	0.10
3	C3, C16, C17	120 pF 0805	311-1112-1-ND	0.10	0.30
1	C15	150 pF 0805	311-1113-1-ND	0.10	0.10
1	C1	270 pF 0805	311-1116-1-ND	0.12	0.12
6	C6, C9, C11, C12, C18, C19	1000 pF 0805	311-1122-1-ND	0.18	1.08
1	C2	3900 pF 0805	311-1132-1-ND	0.10	0.10
1	C22	0.01 μF 0805	311-1136-1-ND	0.10	0.10
1	C23	0.1 μF 0805	311-1140-1-ND	0.10	0.10
1	C4	0.22 μF 0805	587-1287-1-ND	0.12	0.12
1	C5	0.47 μF 0805	587-1290-1-ND	0.12	0.12
5	C7, C8, C20, C24, C25	4.7 μF 0805	587-1782-1-ND	0.24	1.20
1	TC1	3.5-10 pF	490-1996-1-ND	1.20	1.20
1	TC2	7.5-30 pF	490-1994-1-ND	1.20	1.20
1	L1	10 nH 0603	PCD2002CT-ND	0.14	0.14
1	L4	22 nH 0805	495-1850-1-ND	0.79	0.79
1	L3	68 nH 0805	PCD1170CT-ND	0.33	0.33
1	L2	100 nH	PCD1172CT-ND	0.33	0.33
TOTAL COST					**$44.00**

pads, and connections are provided for in-circuit reprogramming of the PIC micro.

Beacon Parameters

Major parameters in the APRS message are user-defined via RS232 by invoking a special "programming" mode of operation. This is accomplished by installing a wire jumper between pins 1 and 6 of the ICSP pads, followed by a power-on reboot. Programming can be done with *HyperTerminal*, but a "utility" program created for this purpose (called *WinBeacon*) is available on the author's website.[2]

The parameters are saved in the PIC nonvolatile EE memory and include these items:

- Transmit frequency (1 kHz steps)
- Transmit repeat time interval (seconds)
- User's call sign and SSID
- Digipeater ("via") call signs and SSIDs (if any, max = 2)
- APRS map symbol (2 characters)
- Comment (32 characters)

All parameters are encoded into a single line of text that is terminated with a carriage return and sent via RS232 to the transmitter. Upon detection of this text, the new parameters are processed and saved in EE memory. The transmitter then switches to the new frequency and generates a "test" transmission for 9 seconds. (An unmodulated signal for 3 seconds, a 1200 Hz modulation tone for 3 seconds and a 2200 Hz modulation tone for 3 seconds.)

Successive programming messages will overwrite the old parameter data and invoke another test transmission, using the new data. To exit the programming mode, the wire jumper is removed and dc power is briefly interrupted to generate another power-on reboot, which causes the transmitter to restart in the regular APRS operating mode (RS232 input = GPS).

Note: It is important to ***not*** connect a GPS receiver to the RS232 input until the programming jumper has been removed and dc power has been interrupted (in that order). Failure to do so will lead to garbled programming data because the first message transmitted by the GPS will be interpreted as a programming message.

The Transmitter Circuit

The transmitter employs an Analog Devices type ADF7012 low-power transmitter IC with about 25 mW output, which drives a Philips type PD85004 RF FET. The FET is operated "approximately" in Class C, to generate 2 W of RF output at 2 meters when powered from 12.6 V dc. The Analog Devices chip employs a fractional-N type PLL circuit, and the PLL loop filter was designed using a free CAD tool available on the Analog Devices website, called *SimPLL*.[3]

Fractional-N PLL circuits are notoriously noisy and the ADF7012 is no exception to this rule. Adjacent channel noise is significant for perhaps 300 kHz on either side of the carrier frequency, (down about 40 dBc), which can cause noise bursts in nearby receivers operating on the same band. The short (and infrequent) APRS transmissions help to mitigate this, but using this design for prolonged transmissions would require attention to this issue.

[Builders should be aware of the FCC Rules requirements for transmitter spurious emissions. In this case Sub Part 97.307 (e) applies:

The mean power of any spurious emission from a station transmitter or external RF power amplifier transmitting on a frequency between 30-225 MHz must be at least 60 dB below the mean power of the fundamental. For a transmitter having a mean power of 25 W or less, the mean power of any spurious emission supplied to the antenna transmission line must not exceed 25 µW and must be at least 40 dB below the mean power of the fundamental emission, but need not be reduced below the power of 10 µW. A transmitter built before April 15, 1977, or first marketed before January 1, 1978, is exempt from this requirement.

As long as the spurious emissions are down 40 dBc as the author indicates, this transmitter would meet the FCC requirements. — *Ed*.]

The ADF7012 chip is programmed and operated with four wires from the PIC 18F442 microcomputer. Four serial messages, each 32 bits long, are sent by the PIC at the start of each transmission. The 12.288 MHz reference crystal is divided by three to yield a phase detector frequency of 4.096 MHz, which allows the microcomputer to easily tune the ADF7012 in steps of exactly 1 kHz. The VCO operates at 4 times the output frequency and is divided down (inside the chip) prior to application to the FET input.

The ADF7012 chip has digital inputs for baseband data modulation of the carrier frequency, but those inputs are not employed in this design. Instead, an 8-bit parallel DAC is employed to generate a modulation sine wave that is applied to the reference crystal with a varicap diode. This is done because 2 meter APRS specifies Bell 202 modulation, which requires audio sine wave modulation of the carrier. The baseband modulation inputs of the ADF7012 chip are strictly digital and do not support sine wave modulation. The modulation characteristics of the varicap diode are nonlinear and the varicap bias resistors have been adjusted (experimentally, with some effort) to achieve the best "balance" between waveform quality, modulation level, and equality of (transmitted) tone levels.

Power for the ADF7012 chip passes through two diodes, to reduce the 5 V supply down to about 3.8 V. While developing this design, my attempts to operate these chips at 5 V yielded several chip failures (data sheet says max = 3.9 V). A 3.6 V regulator was then employed, but I found that the in-circuit programming feature of the PIC chip required at least 4.2 V to work properly. Therefore, the present design uses a 5 V regulator for the benefit of the PIC chip with two diodes in the ADF7012 supply line, for the benefit of the ADF7012 chip.

Provisions are made for applying positive dc bias to the PA stage FET gate during transmission periods (bias = zero at other times). The values of resistors R5 and R6 set the dc bias applied to the FET during transmissions. For 2 meter operation, no dc bias is required, (these resistors are omitted and C12 is shorted), but operation on other bands might require some dc bias to increase the FET gain, by driving it closer to a class B (or even class A) operating point. A trimmer cap in the FET input circuit allows tuning for maximum output power.

The output circuit consists of a Pi network circuit using a circuit board "printed" inductor and fixed capacitors. Output power is typically 2 W when dc power is 12.6 V, but operation at lower voltages is possible. The output power closely follows a square law relationship to dc supply voltage, so cutting the dc supply in half will reduce RF power to about 0.5 W. The voltage regulator for the PIC chip is a garden variety 78L05 that suffers "dropout" at an input of 7 V, so operation below 7 V is not recommended.

NMEA RS232 GPS data is fed to a simple transistor inverter circuit, and then to the UART input of the PIC microcomputer. This input is also used to program the APRS parameters when the PROGRAMMING mode is invoked. The 78L05 regulator uses a 100 mA TO-89 package, with abundant heat sinking to the circuit board. There is sufficient surplus current capacity to provide 5 V dc (if desired) to most GPS "puck" receivers that typically would be used with this transmitter.

The PIC Software

The software is written in PIC assembly code for a PIC type 18F442 microprocessor in a TQFP package, running with a 20 MHz crystal clock. It contains 4000 lines of source code and assembles into 3400 bytes of machine code.

TMR0 is used to generate the Bell 202 modulation tones (1200/2200 Hz). Each sine wave is generated in 63 time steps (one step per TMR0 interrupt) using a sine wave lookup table and an external 8 bit DAC. The time interval between successive steps is controlled by the TMR0 interval, which can take one of two different values: one for a 1200 Hz tone, another for a 2200 Hz tone. The reload value for TMR0 is fetched from a RAM location each time TMR0 generates an interrupt, so changing the value stored in this RAM location provides a way to generate a phase contiguous shift between tones.

TMR1 is used to generate the 1200 baud data rate clock for the APRS message bits. The TMR0 reload value (stored in RAM) is modified in the TMR1 interrupt routine, if the next data bit to be sent is a logic "0" bit (logic "0" bits require a tone shift, logic "1" bits do not).

TMR3 is used for the transmit repeat time interval (the time between successive beacon transmissions). TMR3 is (briefly) disabled during transmissions because it (somehow) causes significant interference with the transmitted tones.

The UART (receive only) is used to accept RS232 GPS messages during normal operation, and to accept "parameter programming" messages when operating in PROGRAMMING mode.

Events generated by TMR0, TMR1, TMR3 and the UART RX are all handled in the interrupt routine. The starting address for the interrupt routine is dictated by the internal architecture of the PIC chip, and is (for these chips) equal to 0x08. The CSTART routine (ColdStart, at address 0x0) merely jumps to the main startup routines, to bypass the interrupt routines, which must begin at address 0x08.

The MSTART routine (MainStart) contained the main startup routines that configure the various internal features of the chip, initializes the variables and enables the interrupts. The presence/absence of the PROGRAMMING jumper is also detected in this routine, and if detected, control passes to the HOST_START routine (for changing the APRS parameters via RS232).

The EXEC routine is the top-level routine that supervises operation after completion of the MSTART code. The first part decides if a transmission is required, either a timed transmission or a "forced" transmission, caused by a switch closure to ground on PORTB, bit 6. If a transmission is required, control passes to the EXEC_SEND routine, which then gets fresh data from a GPS message and then calls the SEND_MESSAGE routine to actually generate and send the message.

The GPS message is received and parsed into individual data fields as it arrives, one character at a time, in the UART interrupt routine. The heart of the GPS parsing routine is the RMC_PARSE jump table, which acts much like a SELECT CASE statement found in higher languages. The jump table has 23 possible exit paths, (goto statements) selected by an index variable called GPRMC_STATUS.

As each character (in the GPS message) is detected and parsed in those 23 routines, the GPRMC_STATUS pointer is incremented (when each parsing task is finished) so that the next GPS character that arrives will be "handed off" to the next routine in the parsing sequence. When (and if) the first 22 parsing routines are successfully completed, the last parsing routine (GPS_LF) searches for an ASCII <lf> (line feed) character, which signals the end of the GPS message. This triggers a few more housekeeping chores to adjust the GPS data (if required) and a flag bit is set to indicate that a complete set of valid GPS data is available. This flag bit is tested in the EXEC routine, which waits until it is set before starting an APRS transmission.

Various tests are performed in the parsing routines, to ensure the data is correct. In some cases, the parsing routines will wait for a specific character, and then increment the jump pointer when the character is detected. In other cases, any errors in the GPS data will trigger a restart of the search sequence and the GPRMC_STATUS pointer will be reset to the beginning of the search. Using this method, only a valid GPRMC message will successfully navigate through all 23 parsing routines and cause the flag bit to be set.

A similar jump table scheme is used to generate the APRS message (from a lookup table of characters) in the TEXT_TAB routine located in the SEND_MESSAGE routine.

The actual bytes and bits to be transmitted (once they are fetched) are processed in the SEND_BYTE, SEND BIT and FLIP_TONE routines. The SEND_BYTE routine does the bit stuffing (if required) and calls the CRC_CALC routine for each data bit, to keep the CRC checksum value honest. The TX_START and PLL_SEND routines start up the ADF7012 PLL chip each time a transmission is performed, by sending 4 bytes to each of 4 registers in the ADF7012 chip (16 bytes total).

The HOST routines are only used if the PROGRAM MODE is selected (ground strap was detected when dc power is turned on). The PROGRAM mode is used to set the various parameters of the transmitter, including call sign, repeat time, APRS map symbol, comment and so on. The HOST routines wait for a complete programming message to arrive via RS232, then they parse the message and save the various parameters in nonvolatile EE memory. This is followed by a test transmission.

In-Circuit PIC Programming

For experimenters, provisions are made for reprogramming the PIC chip *in situ* using a simple cable that can be made to mate with a PIC programmer. Six holes (circuit board vias) are located along the top edge of the circuit board near the PIC microprocessor. These holes are spaced on centers of 0.1 inch and are large enough to allow insertion of the 0.025 inch square wire wrap pins, like those typically used on wire wrap IC sockets or garden variety SIP header strips.

In a sense, these holes provide the in-circuit programming socket and the mating strip of wire wrap pins (inserted into these holes) would be the programming plug. A simple cable can be contrived to connect this plug to a PIC programmer to allow in-circuit reprogramming of the PIC chip. (*Hint:* Hold your finger on the plug to maintain good contact between holes and pins while programming the PIC chip. It is a loose, unreliable fit otherwise.)

Five wires are required, but the PIC chip allows two different programming methods (high voltage or "HV" method as well as a "5 V" programming method) so 6 pins are provided to allow user selection of the desired approach. Pin 1 is signified with a square pad; the others are round. The pin definitions are:

- Pin 1 PGM pin (used ONLY for 5 V programming mode)
- Pin 2 PGC pin (programming clock)
- Pin 3 PGD pin (programming data)
- Pin 4 V_{PP} pin (programming pulse, used for HV programming mode)
- Pin 5 V_{DD} pin (+ 5 V dc power)
- Pin 6 GND pin (ground)

For example, using a PicKit 3 programmer, the following cable would work:

PicKit 3	*MicroBeacon*	*Signal Name*
Pin 1	Pin 4	V_{PP} (HV programming pulse)
Pin 2	Pin 5	V_{DD} (5 V supply
Pin 3	Pin 6	GND (ground)
Pin 4	Pin 3	PGD (programming data)
Pin 5	Pin 2	PGC (programming clock)
Pin 6	Pin 1	PGM (5 V programming select)

A PicStart Plus programmer expects an actual IC to be inserted into its ZIF socket, but a cable can be made for external programming by using a wire wrap IC socket as a "dummy" chip, which is inserted into the programmer's ZIF socket. Cable wires running to the beacon programming "plug" can then be soldered to the WW pins of this socket.

Note: To program the PIC chip, the beacon dc power must be turned on. Do not rely on the programmer to provide 5 V dc power. For regular "HV" programming mode, the only connections actually required are V_{PP}, ground, PGC and PGD.

The Circuit Board CAD File

The circuit board CAD file uses the ExpressPCB "MiniBoard" service, and was created with the (free) CAD tools available from ExpressPCB.[4] Their "MiniBoard" service is highly standardized and inflexible, to allow maximum economy, and the CAD file provided here will generate nine beacon PC boards, for a total (bare board) cost (June 2011) of US $65.

The boards are paneled in this design (to comply with MiniBoard requirements) and must be separated from each other using a hack saw, coping saw, or some similar means.

The Parts List

Except where indicated, all part numbers are DigiKey; prices and quantities are for a single beacon transmitter. Parts marked ALT are alternative parts (in case the primary part is not available). Quantities and costs are per transmitter.

Bob Simmons, WB6EYV, was first licensed as a Novice in 1964 at age 13 and remained licensed (more or less) constantly ever since. He also earned a commercial FCC license in 1967. Bob served Naval Reserve duty as a radar technician with about 6 months of total sea time. He spent several years doing civilian work in nautical and marine electronics in Los Angeles harbor, as well as doing some land mobile radio work. This was followed by five years in flight line avionics, working with business jets. He moved to Santa Barbara, California in 1992 and worked on vacuum deposition systems for five years and held assorted engineering jobs at other times. Presently, Bob is self employed and runs a website, making and selling radio direction finding equipment and modules, with the majority of his latest work spent creating embedded software/hardware and developing technologies to enable Internet-linked remote DF stations. His primary interest is developing and applying new technologies to old problems and pushing the DF art forward.

Notes

[1]The ExpressPCB circuit board files and the program code files for this article are available for download from the ARRL *QEX* files website. Go to **www.arrl.org/qexfiles** and look for the file **7x13_Simmons.zip**.

[2]You can download the *WinBeacon* program at: **www.silcom.com/~pelican2/PicoDopp/XDOPP.htm#WBCN**.

[3]The *SimPLL* CAD program is available for download from: **https://form.analog.com/Form_Pages/RFComms/ADISimPll.aspx**.

[4]Learn more about the *ExpressPCB* CAD program and download the files at: **www.expresspcb.com**.

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The HF7070 HF/LF Communications Receiver Prototype

A detailed look at high performance receiver design.

The HF7070 receiver is a double conversion superheterodyne with a first IF at 45 MHz and a second IF that is centered at 44 kHz before going to a 25 bit audio ADC. The output from the ADC goes to an advanced 24 bit fixed-point DSP system. The radio covers dc to 30 MHz and has a noise figure of 12 dB without the use of a preamplifier before the first mixer. Out-of-band IP3 at 50 kHz spacing is 45 dBm, giving an SSB IP3 dynamic range of 115 dB. The IP3 within the 15 kHz bandwidth of the roofing filters is 19 dBm at 100 Hz spacing in an SSB bandwidth, which results in an IP3 dynamic range of 97 dB. This in-band linearity sets new standards for an up-conversion radio and gives superb high fidelity reception of FM, AM, SSB and CW signals.

To complement its excellent technical performance there are all the usual DSP features for the user. These include a sensitive band scope on the LCD panel of the radio. The band scope can also be displayed on a computer connected via a TOSLINK optical cable. With a noise floor of −145 dBm in a 50 Hz bandwidth, it can display submicrovolt signals.

The receiver analog front end has two H-Mode mixers using fast bus switches. The first mixer is terminated by quadrature hybrid-connected two-pole 45 MHz filters of 15 kHz bandwidth, which is followed by the first IF amplifier. This amplifier, with a noise figure of only 1.3 dB and an IP3 at 40 dBm, drives 4 poles of roofing filter, which is followed by a second amplifier. This amplifier drives the second H-Mode mixer that gives an output centered on 44 kHz. There is a balanced stage of amplification at 44 kHz before the 25 bit audio ADC. The 6 poles of roofing filter at 45 MHz gives 115 dB image rejection at the second H-Mode mixer, so an image rejection mixer is not required.

The HF7070 was designed by the British electronic engineer John Thorpe of JTdesign based in Matlock, England. John also designed the Lowe receivers and the highly acclaimed AR7030 HF/LF receiver manufactured by AOR UK.

In terms of its technical performance the AR7030 represented very good value for money and was made from 1996 until production ceased in 2007 due to the restriction of hazardous substance (ROHS) directive. Over that period some 5000 units were sold, most of which went for export (even the French Navy bought a few).

Originally, John designed the HF7070 receiver for AOR UK, but they ceased trading two years ago. John is a consultant and had other design commitments. So, I designed and built an up-conversion front end for the 7070 a few years ago to help with the development work. John was able to re-engineer this for mainly surface mount components and improve on its technical performance.

Just after John was presented with the front end board, I got to know Martein Bakker, PA3AKE. Martein was keen to build a holy grail version of the CDG2000 transceiver and had made technical measurements on transformers and fast bus switches for H-Mode mixers. These measurements proved quite useful to John. Those readers who are familiar with the CDG2000 Amateur Radio transceiver project will recognize the similarity of the HF7070 front end block diagram (see Figure 1) to the receiver in the CDG2000. Any reader who is interested in the detailed technical measurements for the HF7070 "proto2" receiver will find them on PA3AKE's website at **http://martein.home.xs4all.nl/pa3ake/hmode/**.

Fundamental Design Issues in the Analog Signal Path

Table 1 shows some of the key parameters of the receiver front end design. It is the job of the two-pole 45 MHz filter to provide some protection to the first IF amplifier for strong signals more than 10 kHz from the selected frequency.

In an up-conversion radio the third order intercept (IP3) usually reduces significantly for off-channel signals within the bandwidth of its roofing filters. The close-in performance of the HF7070 has surprised quite a few people and Table 1 shows how this excellent close-in IP3 performance is achieved. The values in the table of Net NF and Net IP3 are slightly better than the practical measurements on the "proto 2" receiver. In the table, the noise figure (NF) of the receiver at the antenna is 11 dB. This gives a noise floor of –129 dBm in a 2.4 kHz bandwidth. Together with a Net IP3 of 24.5 dBm the result is an in-band dynamic range of 102 dB. The practical results are NF 12 dB and an IP3 of 19 dBm, giving an in-band IP3 dynamic range of 97 dB.

It is necessary for the analog front end to have enough gain so that noise from the analog stages dominates the quantization noise from the ADC in the digital signal fed to the DSP. This gives a smooth, audible transition from noise to signal.

Referring to Table 1, in-band IP3 is limited by the 4-pole roofing filter. As the net IP3 requirement increases, as the signal is amplified the linearity will be degraded if the net value exceeds the stage IP3. This does not happen with the HF7070, but it is interesting to note that the 44 kHz amplifier has an IP3 output of 60 dBm.

A paradox is that a bit of signal path loss at the right place is a design virtue. The IP3 dynamic range at a 100 Hz test tone spacing in a 2.4 kHz bandwidth is 97 dB. The Yaesu FTDX-5000 has an IP3 of –8.5 dBm within the bandwidth of its 9 MHz roofing filters and a NF of 17 dB, which yields a dynamic range of 77 dB. So you can see why people with knowledge of receiver design are impressed with the close-in performance of the HF7070.

The First H-Mode Mixer

The experimental 7070 front end that I gave to John used the Fairchild FST3125 fast bus switch in the mixer. An important result from Martein's experimental work was that the more recent Fairchild FSA3157 actually proved to be the best switch for use in H-Mode mixers. This is a SPDT switch with a 0.5 ns break-before-make action. This SPDT switch is ideal for being driven by a fundamental frequency squarer because it is not necessary for the drive logic to generate a complement signal, as was the case with the

Table 1
Key Parameters of the HF7070 Front End

	Stage Gain dB	*Net Gain dB*	*Stage IP3 dBm*	*Net IP3 dBm*	*Stage NF dB*	*Net NF dB*
Antenna low pass filter	–2	–2	45	24.5	2	11.1
First H-Mode mixer	–5	–7	45	22.5	5	9.1
2 pole 45 MHz filter	–1.5	–8.5	24	16	1.5	4.1
First 45 MHz amp	10	1.5	40	26	1.3	2.6
4 pole 45 MHz filter	–2	–0.5	24	24	2	7.7
Second 45 MHz amp	10	9.5	40	34	1.3	5.7
Second H-Mode mixer	–5	4.5	45	29	8	14.02
44 kHz amp	18	22.5	60	47	6	6.02
25 Bit ADC	0	22.5	50	47	4	4

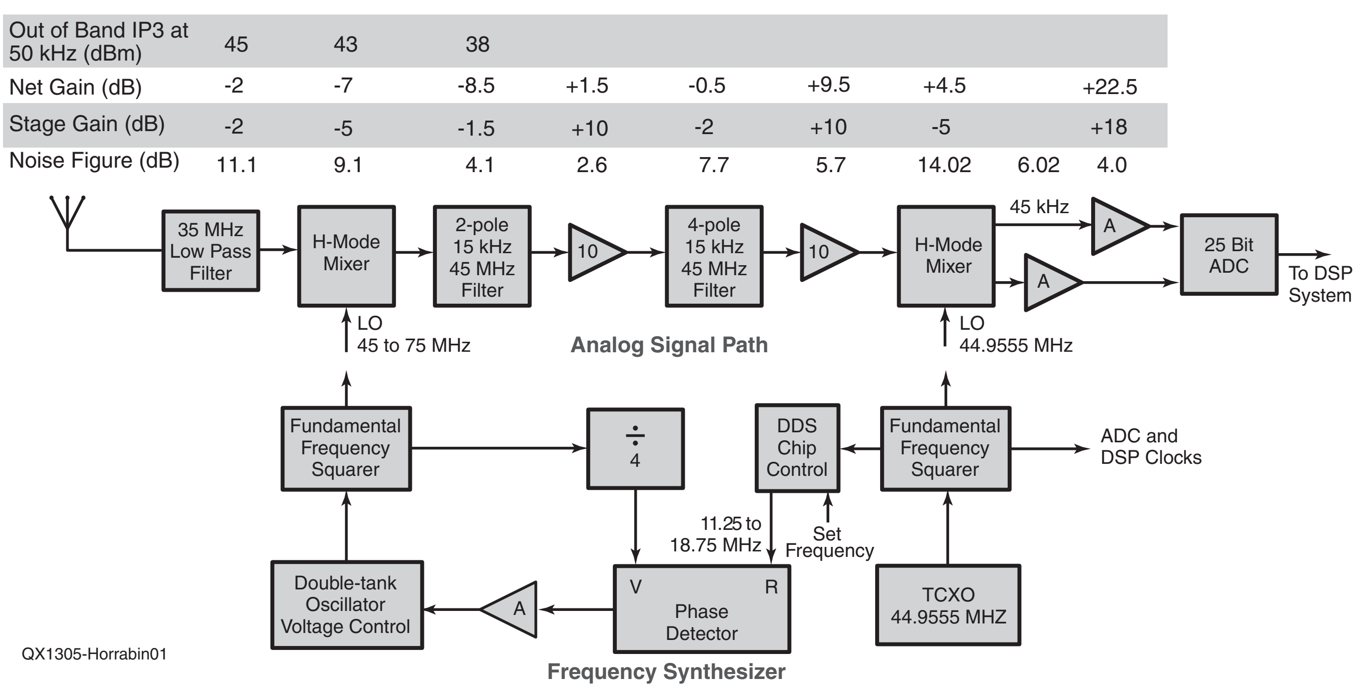

Figure 1 — The HF7070 front end block diagram with Out of Band IP3, Net Gain, State Gain and Noise figure indicated for various stages.

FST3125. The choice of Mini-Circuits transformers used in this mixer gives the radio an out-of-band IP3 of 45 dBm at a 50 kHz test tone spacing with high sensitivity down to an input frequency of 10 kHz.

15 kHz Bandwidth 2-Pole 45 MHz Crystal Filter

Two of these filters are connected via quadrature hybrids to terminate the mixer. The use of quadrature hybrids with two identical filters always presents a nominal 50 Ω termination to the mixer, even when individual filters present a reactive load.

A design goal of the HF7070 was to build a sensitive receiver without a preamplifier before the first H-Mode mixer. Because the HF7070 is a general coverage multimode receiver, it was always our intention to use a 15 kHz bandwidth roofing filter system to accommodate FM and DRM signals. Another reason for this is that narrower bandwidth crystal filters have lower design impedances, which would give a greater insertion loss and therefore require a preamplifier before the first mixer. This filter and its surrounding circuitry introduce a loss of only 1.5 dB.

The First 45 MHz Amplifier

This amplifier drives the second roofing filter and it is the input IP3 of this filter (26 dBm) that limits the in-band performance of the radio.

The amplifier is a 45 MHz version of the 4 × J310 amplifier designed by Bill Carver, W7AAZ, as used in the CDG2000 transceiver. The noise figure of this amplifier is 1.3 dB. Its output IP3 of 40 dBm makes it a particularly important building block in the HF7070 to satisfy the requirement of having a sensitive, linear receiver without a preamplifier before the first H-Mode mixer. This amplifier uses source gate feedback, which gives excellent reverse isolation so that its input impedance is not affected by its output driving a crystal filter, which in its transition region can present a reactive load.

15 kHz Bandwidth 4-Pole 45 MHz Crystal Filter

This filter has an in-band IP3 at its input of 26 dBm and this ultimately controls the in-band IP3 of the radio. Some Amateur Radio equipment manufacturers offer narrower bandwidth VHF roofing filters for their radios. You don't want this for two reasons: (1) the insertion loss increases and (2) the in-band IP3 for a narrower design bandwidth with the same quality of quartz crystals gets worse at roughly 6 dB per octave. This effect was discovered by PA3AKE when he was designing his 9 MHz roofing filters with the fabulously linear crystals supplied by the German firm Quarztechnik.

So what you really need in an up-conversion radio is a wide roofing filter, but the linearity of following circuitry must not degrade the in-band IP3 of the crystal filter. This seems to be a problem area with many transceiver designs.

The second 45 MHz amplifier is similar to the first amplifier and uses the 4 × J310. Its output IP3 is 6 dB higher than the net IP3 at that point.

The second H-Mode mixer gives 115 dB rejection of the image because of the stop band of 6 poles of roofing filter, so an image rejection mixer is not required. This mixer gives a push-pull output centered on 44 kHz.

The 44 kHz Amplifier

John designed this amplifier with a noise figure of 6 dB, a gain of 18 dB and an output IP3 at 60 dBm. A point worth noting is that it is easier to get low-noise, high-IP3 gain at 45 MHz than at audio frequencies. This was an important consideration for the gain distribution of the front end. There is also a potential problem for the unwary that may affect other designs: many operational amplifiers have poor IP3 characteristics.

You get good IP3 results at high signal levels, but the IP3 tones do not fall away as the signal amplitude is reduced; they remain at 60 dB down. There were a few operational amplifiers that gave the correct IP3 behavior.

The reason for this problem may be due to crossover distortion in the output stage of most operational amplifiers. John was aware of this problem when he designed the 44 kHz amplifier. It is interesting to note that the new ICOM IC-7410 with its 36 kHz IF appears to use a version of the TL074 that is known to have the IP3 problem.

Apart from the front panel board, the radio has an analog board and a digital board. After some discussion the 25 bit ADC went on the digital board. The second H-Mode mixer makes the transition from 45 MHz to 44 kHz and it gives a push-pull balanced output. This is followed by a balanced amplifier at 44 kHz located on the analog board whose outputs are connected by a short length of strip cable from the analog board to the balanced inputs of the 25 bit ADC located on the digital board. Whatever common mode noise is present due to the strip cable connection between the two boards is well within the common mode rejection of the ADC.

The 25 Bit ADC

The radio uses a top-of-the-range, stereo, 24 bit audio ADC with both channels driven and the signals digitally added giving a theoretical 25 bit performance, It is worth remembering that the ADC is effectively the main gain block. An estimate of its noise figure

Further Developments

The manufacture of the ten HF7070 production prototypes has been funded by a British company owned by a radio amateur. He was happy to learn that this article was appearing in *QEX*, but he asked that neither he nor his company be mentioned. In view of this it seems unlikely there will be a significant production run because he is concerned that there would be too much competition from SDRs.

Collaboration between British and American electronic engineers during World War II resulted in some great design work. The thing to remember about the HF7070 is that although it has been designed by a British electronic engineer, all the components that really matter, apart from the 45 MHz monolithic filters with their excellent IP3s, were designed in the USA. At least in that sense, history is repeating itself.

It would have been more cost effective to use an MMIC in the 45 MHz IF strip, but nothing was available that met the technical requirements. That has changed recently, however. The American firm Mini-Circuits has introduced the PHA-1+ and the dual-matched version known as the PHA-11+. Although they were designed for microwave applications, the noise figure (NF) and IP3 out is at its best between 45 MHz and 100 MHz, making them useful in IF amplifiers for up-conversion radios. Typical performance at 45 MHz is: NF 2.2 dB: gain 18 dB: and IP3 out 42 dBm.

The vector network analyzer designed by Paul Kiciak, N2PK, together with the *Windows* software for the VNA written by Dave Robert, G8KBB, was used to measure the PHA-1 input impedance at 45 MHz, which is 80 Ω in parallel with 25 pF. The best way to use this part in the existing HF7070 IF strip is to use two in parallel with an output attenuator to replace the first 4 × J310 amplifier, and one to replace the second amplifier, again with an output attenuator.

A change to the HF7070 front end architecture by having all 6 poles of roofing filter connected via quadrature hybrids immediately after the first H-Mode mixer could, in principle, further increase close-in receiver dynamic range. Obviously, careful shielding around the filters is required. All the mechanical and electronic components for this experiment have been obtained and I hope to report the results in a future *QEX* article. From a practical printed circuit point of view, rather than use the matched pair the thermal affect on amplifier noise figure would have been reduced if two individual PHA-1s could have been used. Unfortunately, this did not fit well with the circuit board design. Another version of the PHA-1 where the input and output pins are transposed could be useful. Perhaps Mini-Circuits will consider this.

(NF) is 4 dB, but if it was 12 dB the effect on receiver noise figure would be to increase it by only 0.1 dB.

Overall Front End Performance of the Signal Path

The HF7070 has excellent technical specifications and this is reflected in its outstanding on-air performance. This didn't happen by chance; a lot of thought has gone into the receiver front end architecture.

It is unusual to have a sensitive up-conversion receiver that does not use a preamplifier before the first mixer. In many ways the front end architecture makes use of the principles established by Bill Squires, W2PUL, in the 1960s in his SS-1R receiver. His receiver used the 7360 beam-switching tube in the mixers and, although it was a tunable IF design, gain distribution was carefully controlled before the main crystal filter. It also lacked a preamplifier before its first mixer.

Just like PA3AKE's holy grail receiver with a 9 MHz IF, the HF7070 is only as good as the in-band and out-of-band linearity of its crystal roofing filters. Martein has made IP3 measurements on a number of different manufacturers' 15 kHz bandwidth, 45 MHz fundamental-mode monolithic crystal filters. Only one makes the grade in terms of IP3 and its 3rd order law compliance with input signal level.

A Low Phase Noise Local Oscillator for the HF7070

The local oscillator in the HF7070 is a DDS/PLL design using a double-tank VCO. This was designed by John Thorpe and it is similar to the one used in the AR7030 receiver, but with improved performance. The basic double-tank oscillator is shown in Figure 2. The principle involved is that this circuit can only oscillate if the cold end of the active tank (the one with the J310) is a low impedance to ground. This can only occur if the dummy tank is series resonant at the same frequency as the active tank. This means that as you move away from the carrier two resonators are active, which increases the rate of phase noise fall-off with offset frequency. This circuit has never been analyzed from a phase noise point of view by a mathematician, but measurements show phase noise falls off at 30 dB/decade compared to 20 dB/decade in a single resonator oscillator.

In a superhet receiver local oscillator, sideband noise causes reciprocal mixing and, as a result the dynamic range associated with good close-in IP3, can be limited by the sideband noise of the local oscillator. The frequency reference used in the HF7070 frequency synthesizer is a 44.9555 MHz TCXO that has a good phase noise profile. The TCXO is used for the local oscillator feed to the second H-Mode mixer and it also clocks an AD9951 DDS chip. Other divided outputs from the TCXO are used to provide clocks for the 25 bit ADC and the DSP system.

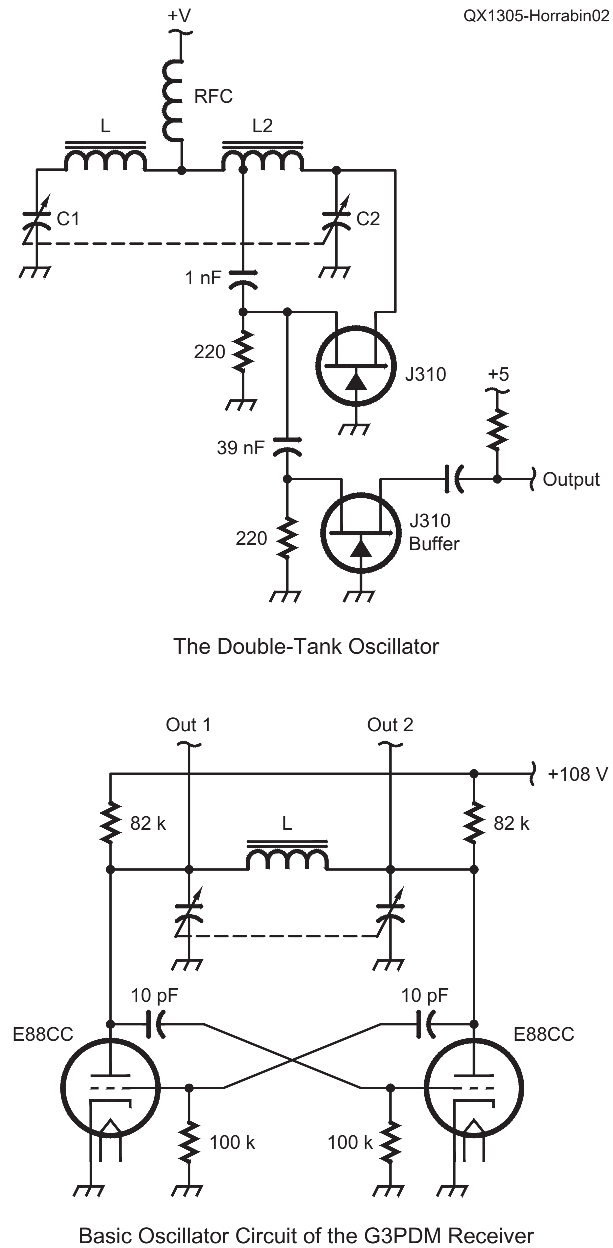

Figure 2 — The HF7070 double-tank oscillator (top) compared to the tube-based oscillator used in the G3PDM/W1 receiver designed by Peter Martin in 1970.

Operation of the local oscillator frequency synthesizer is as follows. The output frequency from the double-tank VCO is divided by 4, which is then phase locked to a DDS-generated reference frequency in the range of 11.25 to 18.75 MHz. The result is that the VCO tunes 45 to 75 MHz in small steps. The VCO is buffered and goes to a fundamental frequency squarer whose output drives the FSA3157 switches in the first H-Mode mixer. The loop bandwidth of the PLL system is about 1 kHz, within which VCO oscillator phase noise is reduced by the action of the loop.

A measured value of phase noise for the HF7070 at 200 Hz to 1 kHz offset on the 7 MHz band is a −116 dBc/Hz plateau falling to −128 dBc/Hz at 7 kHz. Six SSB signals could be present within the 15 kHz bandwidth of the roofing filter and would be separated in at least an 82 dB dynamic range because of reciprocal mixing

For CW reception with a 250 Hz DSP, the bandwidth dynamic range would be at least

Table 2
Phase Noise Comparison

Frequency Offset (kHz)	*3*	*5*	*10*	*20*	*30*	*50*	*100*	*200*
AR7030 dBc/Hz	–113	–118	–128	–137	–142	–147	–155	–160
HF7070 dBc/Hz	–116	–122	–133	–142	–145	–150	–158	–162
HF7070 at 7 MHz	–120	–126	–138	–147	–150	–154	–159	–162

92 dB. If the DSP bandwidth was reduced to 50 Hz you could resolve CW signals in at least a 99 dB range.

Outside the PLL bandwidth the phase noise profile is that of the double-tank oscillator itself. The HF7070 is not phase noise limited beyond 7 kHz from the carrier, reaching –150 dBc/Hz at 30 kHz from the carrier on the 7 MHz band. According to Leeson, phase noise in a single-tank oscillator reduces beyond a corner frequency at 20 dB per decade as you move away from the carrier. In the double-tank oscillator our measurements have shown that phase noise decreases by about 30 dB per decade.

By examining Table 2 you'll see that phase noise falls at around 30 dB per decade of frequency offset in both the AR7030 and the HF7070. The results for the AR7030 are calculated from the Peter Hart review of the AR7030 receiver in the July 1996 *RadCom*.

The results characterize the basic phase-noise profile of the AR7030s double-tank VCO, which is the result of the very narrow PLL bandwidth used in the AR7030 receiver. The HF7070 is added for comparison and adjusted to compare with the AR7030 measurements at 21 MHz.

I developed the double-tank circuit in 1994 to complement my design of the H-Mode mixer and John developed it further for use in the AR7030 receiver. The owner's manual for the AR7030 includes a complete circuit of the radio, so it is surprising that John's version of the double-tank oscillator has not been copied for use in other commercial developments (such as other Amateur Radio equipment). In fact, the real origin of the double-tank circuit goes back a bit further than 1994, all the way to the G3PDM receiver designed by Peter Martin in 1970. His receiver used a 7360 as the mixer and he had a push-pull double-triode local oscillator VCO to drive it. There was only one resonator, each end of which went to the plate of a triode. Each triode plate was connected to the high-voltage power supply through an 82 kΩ resistor. You can't do that with J310s, but if you compare the two circuits they do a similar job. So, it is fair to say that Peter got there first.

Semiconductor designers appear to have a love affair with band-gap voltage references in what they consider to be low noise systems. These usually have a voltage of

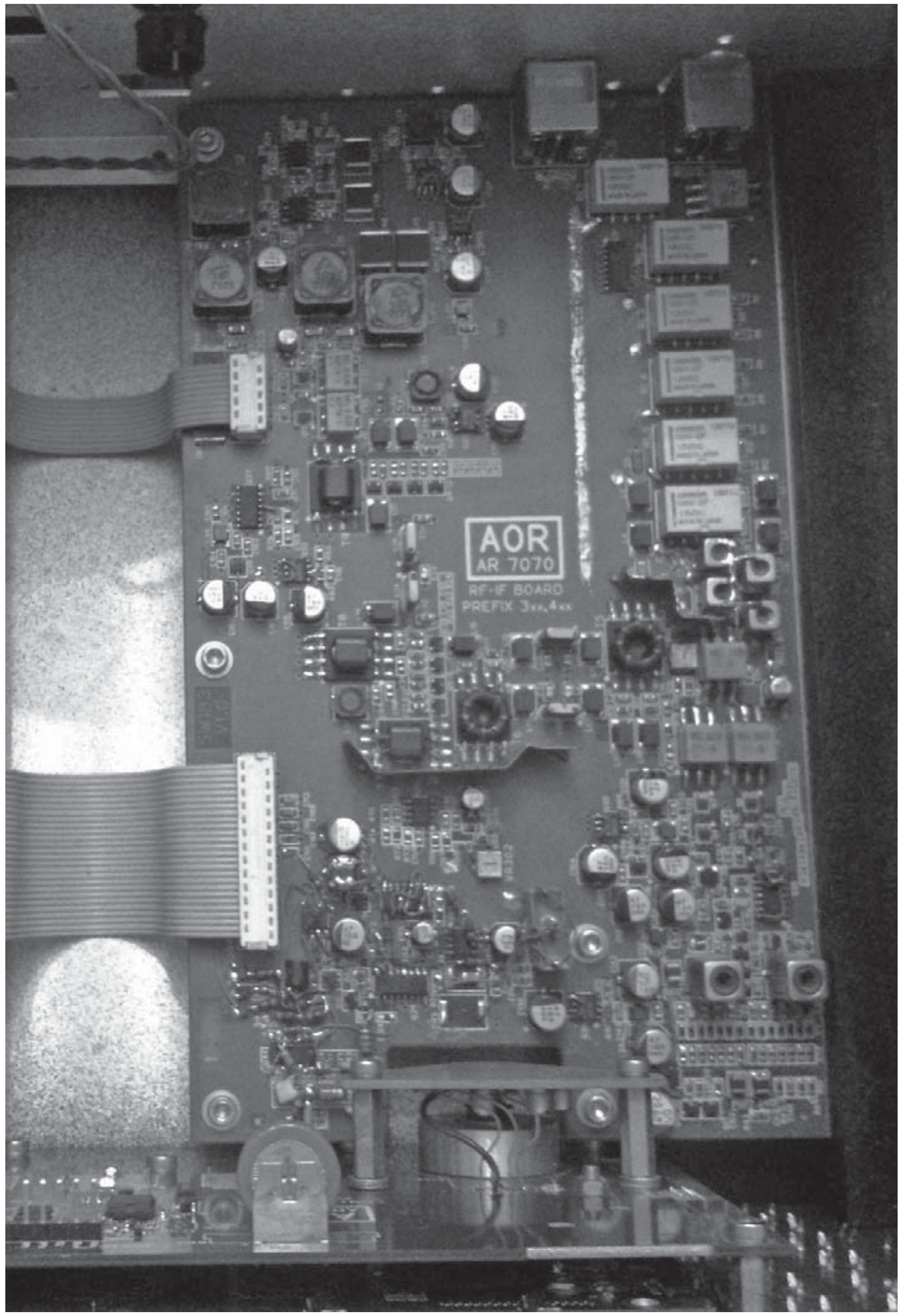

Figure 3 — This photo shows the RF/IF circuit board in the prototype HF7070 receiver. The signals enter at the top right, and the white blocks in that section of the board are relays for an input attenuator. Below the relays are the low pass filter inductors and then the Mini-Circuits transformers of the first H-mode mixer. To the left is the first toroid of the quadrature hybrid, two poles of filtering and the second quadrature hybrid. To the left of center is the 4 × J310 amplifier. In the top left quarter of the board there is a four pole filter, the second 4 × J310 amplifier, the second H-mode mixer and the 44 kHz push-pull amplifier. The narrow cable strip carries the amplified signal to the analog-to-digital converter on the digital circuit board. The bottom section of the circuit board includes the frequency synthesizer and low-noise voltage regulators.

1.2 V and an RMS noise of about 10 μV in a 10 kHz bandwidth. They have a characteristic noise profile with a noise plateau that extends out to about 300 kHz before it falls sharply to the thermal noise floor. Most chip designs that make use of band-gap references fail to provide a pin to which a large capacitor can be connected to reduce low frequency noise from the reference.

There is suspicion that close-in phase noise within the PLL bandwidth on the HF 7070 is limited by band-gap noise on the phase detector chip made by Philips: the 74HCT9046. To improve this situation it is necessary to know how band-gap noise on the chip may be turned into noise that would affect the charge pump. This turned out to be easier said than done.

In the old days manufacturers of integrated circuits would show internal circuits on their data sheets. Philips was contacted to see if they would provide a circuit of the chip so it could be established exactly what effect band-gap noise could have on the charge pump, but a circuit was not forthcoming. They were again contacted, but finally told us they were not interested in pursuing the issue further because no other '9046 users had asked these questions before and the chip was selling well. As far as Philips was concerned, this was a non-issue.

In truth, these issues have only recently arisen because state of the art up-conversion HF receivers are approaching 120 dB dynamic range, and in the case of PA3AKE's holy grail down-conversion receiver, *exceeding* 120 dB. Semiconductor designers really need to take notice of the shortcomings of band-gap references in ultra low-noise systems and provide a pin to deal with low frequency noise from the reference.

On-Air Performance of the HF7070 Receiver

The technical performance of this radio is what Pat Hawker, MBE, G3VA, (SK) would call "superlinear." Not only does the radio have an SSB IP3 dynamic range of 115 dB at 50 kHz, its design answers the problem of how to get good IP3 dynamic range in an up-conversion radio for signals within the bandwidth of its 45 MHz roofing filter (effectively achieving a dynamic range of 97 dB in an SSB bandwidth). What was unknown was how a radio with this built-in in-band linearity would actually sound on the air.

The answer is "amazing." The fidelity is remarkable and what is particularly noticeable is the way selected signals seem to stand out above the noise floor and adjacent signals. The first time you hear this radio you know you are listening to something special. Unless you hear it yourself, the performance is impossible to describe.

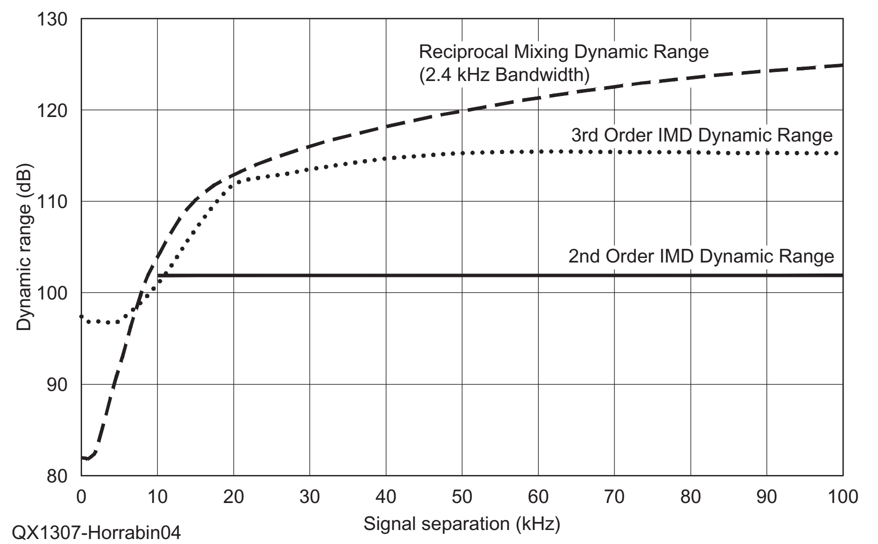

Figure 4 — This graph shows the measured dynamic range (DR) of the HF7070 prototype receiver. The reciprocal mixing dynamic range was measured with a 2.4 kHz bandwidth. Notice that the 2nd order IMD DR is completely flat across the entire measured range. The 3rd order IMD DR increases rapidly above a signal spacing of about 5 kHz and then becomes quite flat at a signal spacing of 20 kHz. Ideally, the IMD DR should not change as the signal spacing of the input signals changes. This radio has a 3rd order IMD DR that is better than 110 dB at 20 kHz spacing, and levels off at 115 dB, which is quite remarkable. This graph was copied from Martein (PA3AKE) Bakker's website: http://martein.home.xs4all.nl/pa3ake/hmode/ar7070-proto2.html.

How I Came to be Involved in the Design of the HF7070

In 1995 I had suggested to John Thorpe that he consider the double-tank oscillator for use in the AR7030 receiver. We had spoken many times on the phone, but we didn't meet until 1998.

In 2000 Dave Roberts, G8KBB, George Fare, G3OGQ, and I started to design the CDG2000 transceiver that used an H-Mode mixer with the FST3125 fast-bus switch. As an analog radio, it made use of the 4 × J310 amplifier and the 9 MHz IF amplifier design that Bill Carver, W7AAZ, presented in *QST*.

When we were discussing the frequency synthesizer, I suggested that we ask John and AOR UK if they would let us use John's VCO design. They did. Since the CDG2000 receiver section uses a 9 MHz IF, the output of John's VCO was divided by at least two, giving –150 dBc/Hz at 25 kHz on 20 meters and –150 dBc/Hz at 12.5 kHz on 40 meters.

The divider was a 74AC74 bistable, so it produced a square wave output, giving both not-Q and Q signals to drive the mixer, and required no adjustments to give a receiver IP3 of 40 dBm.

On one of my visits to John I stopped by AOR UK to thank them for their help. While I was there, Mark Sumner told me that John was investigating a DSP based receiver then known as the AR7070, but progress had been slow. Originally John intended the signal path to be 45 MHz/455 kHz/20 kHz to the 24 bit audio ADC.

Mark supplied me with some of the Hertz Technology 15 kHz bandwidth, 45 MHz fundamental mode crystal filters to test. Their transmission characteristics were very good with low insertion loss and a good stop band. I realized that if 6 poles of filter were used you could go straight down from 45 MHz to 44 kHz, simplifying John's original design and reducing the second H-Mode mixer image by 110 dB. Also the very good in-band IP3 of these filters offered the promise of hitherto unobtainable close-in dynamic range. I designed and built the prototype up-conversion front end and gave the design information to John along with the prototype board.

One of my hallmarks at Daresbury Laboratory was the tendency to push any given technology to the limit. That is why I am now experimenting with the use of the Mini-Circuits PHA-1 MMIC, which in principle could put 10 dB on the in-band IP3 and give 45 dBm at 20 kHz in a front end like that used in the HF7070. I don't really see John making use of this development because it is unlikely he will design another up-conversion receiver. Together with a local oscillator using the AD9910 DDS chip with a 1 GHz low-phase-noise clock, however, this development would represent the ultimate up-conversion receiver technology.

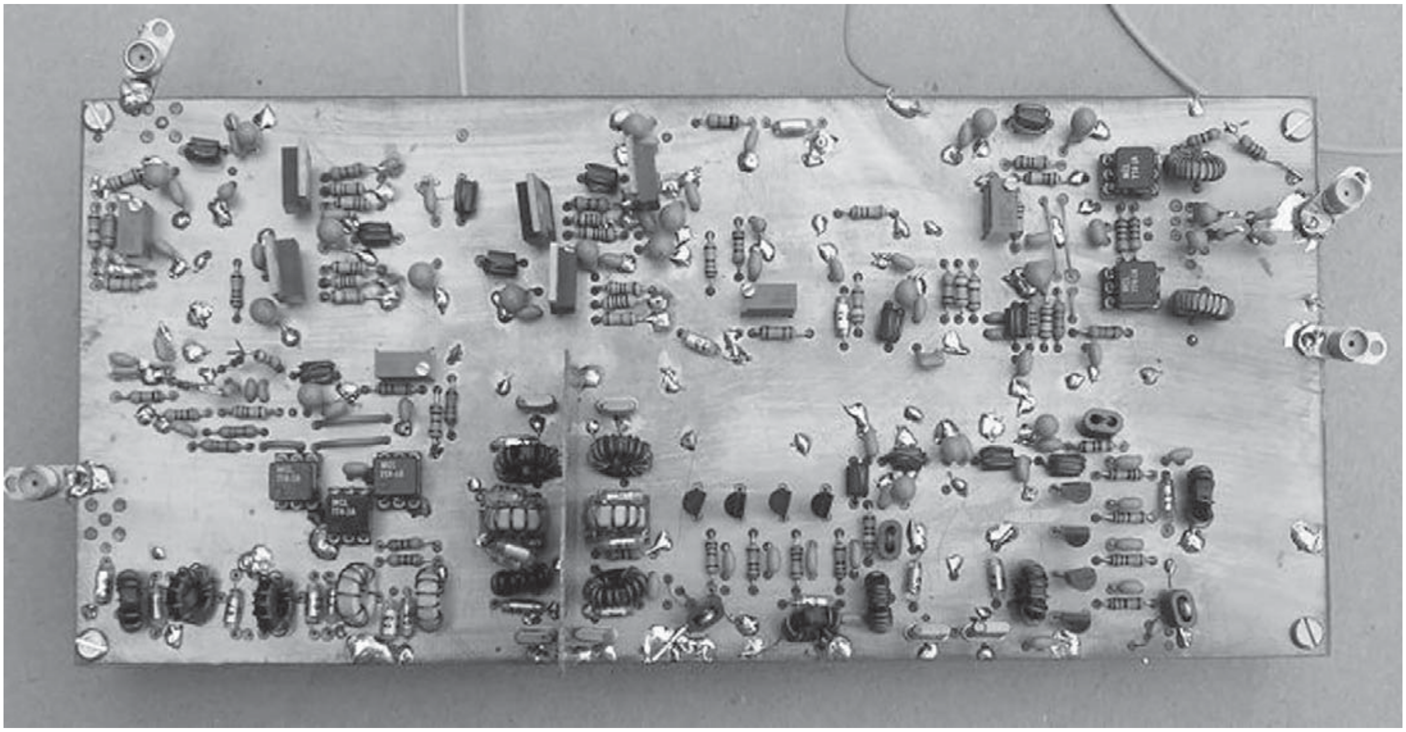

Figure 5 — This photo shows the first prototype receiver front end. It proved the author's design before it was adapted for the HF7070 receiver. This photo is courtesy of Martein Bakker, PA3AKE: http://martein.home.xs4all.nl/pa3ake/hmode/g3sbi_intro.html.

In the 15 kHz wide roofing filter, the signal you are listening to is in the center of the pass band where its group delay characteristics are at their best. This is one factor in the exemplary performance of the radio.

John knows I am a CW enthusiast so my personal agenda was to make sure that the HF7070 was a good CW receiver, and it is. When I first had the opportunity to listen to the HF7070 proto1 receiver, John had a lap top computer connected to display the bandscope. Apart from the larger display, it updates much faster than the LCD display on the radio front panel and I could read CW off the screen. I was watching a DX station on 15 meters working split, and I could see other stations replying. This is remarkable stuff !

Acknowledgements

Because the first part of this article concerns hardware, this is probably the right place to thank all the people who helped design and build the prototype up-conversion front end.

A chance conversation with Pat Hawker, MBE, G3VA, in 1993 about an item in *RadCom's* "Technical Topics" introduced me to the pioneering work of Jacob Makhinson, N6NWP, in receiver front-end design. That set in motion a chain of events that led to my development of the H-Mode mixer and the double-tank oscillator. This

Useful Reading Material

• "Super-Linear HF Receiver Front Ends," (Colin Horrabin, G3SBI, evaluates the N6NWP high performance mixer), *RadCom,* "Technical Topics," Sep 1993, pp 54-56.

• "G3SBI's High Performance Mixer," (H-Mode mixer), *RadCom,* "Technical Topics," Oct 1993, pp 55-56.

• "Crystal Filters for High Performance Mixers," (H-Mode mixer termination with crystal filters using quadrature hybrids), *RadCom,* "Technical Topics," Jan 1994, 37-39.

• Reed Fisher, W2CQH, "Twisted-Wire Quadrature Hybrid Directional Couplers," *QST* Jan 1978, pp 21 to 23. More consistent results can be obtained by using bonded bifilar wire to build your own quadrature hybrids. Such wire is available from the Scientific Wire Company (**www.wires.co.uk/acatalog/bb_wire.html**).

• The CDG2000 transceiver description at the Warrington Amateur Radio Club website at **www.warc.org.uk/proj_cdg2000.php**.

• The G3PDM receiver, as described in the *Radio Communication Handbook* Fifth Edition, 1982, pp 4.50-4.57.

• Wes Hayward, W7ZOI, Rick Campbell, KK7B, and Bob Larkin, W7PUA, *Experimental Methods in RF Design* ARRL, 2009. ARRL Order No. 9239; $49.95. ARRL publications are available from your local ARRL dealer, or from the ARRL Bookstore. Telephone toll free in the US 888-277-5289 or call 860-594-0355, fax 860-594-0303; **www.arrl.org/shop**; **pubsales@arrl.org**.

• Wes Hayward, W7ZOI, and Doug DeMaw, W1FB, (SK), *Solid State Design for the Radio Amateur*, ARRL, 1986. This book is out of print, but it occasionally shows up at Hamfests, on eBay and other used book sites.

• William Sabin and Edgar Schoenike, *Single Sideband Systems and Circuits*, 2nd Ed, McGraw Hill, 1995.

• Peter Hart, G3SJX, "Review — The AR7030 VLF to HF Receiver," *RadCom*, July 1996, pp 44-46.

has led to the high performance receivers found in PA3AKE's holy grail version of the CDG2000 and now the up-conversion HF7070 receiver designed by John Thorpe.

"Technical Topics" also brought together Bill Carver, W7AAZ, Harold Johnson, W4ZCB and Gian Moda, I7SWX. Their contributions to radio technology have been used in the design of the HF7070 as well. The influence of Wes Hayward, W7ZOI, expressed in his various books and articles in *QST* and *QEX* have contributed significantly to the detailed design of the signal path. Research work by Martein Bakker, PA3AKE, presented on his website has been important for the technical status of our hobby and has helped John Thorpe to firm up the choices of some critical components.

I retired in 2000 and it would not have been possible to research and build the prototype up-conversion front end were it not for the good will of former colleagues at Daresbury Laboratory. George Fare, G3OGQ, was recruited to help with circuit board design work and Mark Sumner, G7KNY, formerly of AOR UK (now at MWS Technical Services), provided components needed for experimental work associated with the front end design.

When Martein Bakker, PA3AKE, became involved in testing various H-Mode mixer configurations (transformers and switches), Mark supplied all the bits. It was a real shame that AOR UK ceased trading before the project reached the production stage. The purpose of my experimental front end design was to establish the circuit techniques that John Thorpe was then able to improve upon for his commercial design.

Colin Horrabin, G3SBI, was born in 1941. His father provided him with a World War II BC348 radio receiver for his 12th birthday, followed by a copy of the ARRL Handbook *for Christmas. After years building various projects using government surplus equipment, he obtained his Amateur Radio license in 1963. He has a degree in electrical engineering and a degree equivalent qualification in mechanical engineering. Following an apprenticeship with the British Aircraft Corporation in the early 1960s, he spent over 30 years working at Daresbury Laboratory as an electronic engineer. Colin is interested in small DX antennas for the LF bands, and intends to do some work on small multi turn spiral wound loops that are self resonant containing ¼ wavelengths of wire, which are suitable for transmitting.*

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Tech Notes

The ARRL and its membership supports the recruitment of new Amateur Radio operators and including them in the activity of home brewing, where they design and make their own equipment.

It takes a certain amount of self-confidence and an experimental spirit to get involved in the process of home brewing. In general, one major deterrent to this step is the complexity and perceived difficulty of making the chassis and shielding for one's transmitter, receiver, or transceiver. This may loom as a larger factor than the process of setting up and wiring the circuit itself.

Some people have responded to this situation by building Amateur Radio kits, where the mechanical work of chassis and shielding has already been done, and they have simplified the remaining steps of wiring and soldering the circuit. Many of us older hams were very active with Amateur Radio kits from Heathkit and other kit manufacturers, and remember them fondly.

In this technical note, I recommend taking the next step beyond the typical kit by establishing a shielded environment for beginners' homebrew projects. These environments can be built by individuals and Amateur Radio clubs for the encouragement of new builders.

The Shielded Environment

The basic structure is a box made of conducting metal such as aluminum. This box can be assembled from aluminum sheets bolted together or a commercial aluminum box can be used. As is described in various ARRL references, the bolts should be spaced quite closely to each other so that effective radio frequency (RF) shielding will be achieved. The box could also be made of copper, which will look great, but you will find that copper is quite expensive.

The box is equipped with a door that can be opened so that the novice builder can install and remove his or her circuit boards. This door is intended for occasional use and the metal of the door should be in firm contact with the metal of the box itself so that the RF shielding remains intact.

My recommendation for the size of the box is 24 × 12 × 9 inches. The size of the door should be at least 4.5 × 6 inches to accommodate the installation and removal of circuit boards and prototyping boards, like those sold by vendors such as RadioShack.

Mounting the Circuit Boards

A rack is made and mounted within the box and is fully accessible from the door. This rack is a wood or plastic enclosure that has a set of slots along its opposing sides. The operation of this rack is like the old record cabinets where you slid the records into the slots, which held the records by their edges. In this case, the ham slides the circuit boards into the slots on the sides of the rack.

Each slot is spaced at least one inch from the next slot. This allows a variety of circuit boards to be accommodated in the rack including the following:

- Point-to-point wiring on a perforated board.
- Through-hole mounting on a copper-foil-based circuit board.
- Surface mounting on a copper-foil-based circuit board.
- "Dead bug" wiring over a copper-foil-covered board.

One-inch board spacing should accommodate most wiring situations other than vacuum tubes and/or larger transformers.

Connecting the Boards and Other Items

Wires are used to connect the boards via point-to-point wiring. All of these wires are connected to the top edge of the boards, which are easily accessible at the top of the rack through the open door. This simple point-to-point wiring approach is inexpensive, and will operate effectively in the Amateur Radio high frequency bands. Most of the uses of the shielded box will involve just a few small circuit boards and do not justify the more elaborate backplane and data bus systems used in the computer industry.

Wires are also used to connect the appropriate boards to inputs such as audio and dc power and outputs such as radio frequency signals to an output filter (see description below).

Inputs

One side of the shielded box is dedicated to inputs such as a power switch, dc power, tuning control, output volume, frequency band selection and other controls. Some of these can be installed on the basic box and additional space can be allowed for the builder to install their own. Each input is suitably designed and bypassed so that RF will not leak out at that point.

Outputs

The primary output is an RF Amateur Radio signal, routed from one of the circuit boards through a connector to a filter enclosed in a shielded compartment within the main box. This filter compartment can be opened by the amateur operator and he or she can build their own band-pass or low-pass filter within this space. They can design their own filter and install the components (such as capacitors, inductors, and resistors) in the available volume. The RF output from the filter is run to a UHF-style connector that leads through the shielding wall to a coaxial cable to the antenna.

Secondary outputs provide information to the operator such as voltages or currents on analog dials or digital output displays. A basic panel display can be provided as part of the original shielded box.

Next Steps

The main purpose of the shielded box is to contain circuit boards made by the builder. Third parties can also create designs or completed circuit boards that can be used in the box, however. Some of these third-party products can be complete kits for specific radios such as crystal sets, QRP transmitters, ham band receivers, short-wave-broadcast-band receivers, student radio astronomy receivers (20 MHz), test equipment, and so on. Other third party products could be circuit boards that are populated with sockets for ICs, battery holders, and solderless connection points for wiring.

All of these circuit boards would support the home brewer by providing an environment where boards for specific functions are available in addition to the boards that the builder creates.

Action

If this type of shielded box catches on, a manufacturer, or even the ARRL may want to consider producing it as a special product that will encourage increased participation in radio home brewing. The shielded boxes could be sold in the ARRL store.

Nickolaus E. Leggett, N3NL, is an Amateur Extra Class licensee and ARRL Diamond Club Member (he has been an ARRL member for over 40 years). He is a retired technical writer who used to write user guides and develop online help systems for packet networks (X.25 Protocol). He is an inventor holding three US patents. One of his patents is for a wireless data bus system for computers and other digital systems. His other two patents are on ground effect machines (hovercraft vehicles):

United States Patent 6,771,935, Wireless Bus August 3, 2004

United States Patent 3,280,929, Ground-Effect Machine October 25, 1966

United States Patent 3,280,930, Ground-Effect Vehicle October 25, 1966

(The text and figures of the patents can be accessed online on Google patents.)

Nickolaus holds a Master of Arts degree in Political Science from the Johns Hopkins University (Baltimore, MD). He also has a Bachelor of Arts degree in Government from Wesleyan University, Middletown, CT.

He has been a licensed Amateur Radio operator since the early 1960s, starting with the Novice license, WN2UEQ. He was the slowest Morse code operator on the air, and used a one-tube transmitter built from a kit (Conar). Refer to the Old Radio column in the July 2001 issue of QST *(pages 102 and 103) for photographs of this type of station.*

Nickolaus has published two articles in QEX *and one in* QST*: "The Morphological Table — An Invention Generator,"* QEX*, ARRL, Dec 1987, pp 12-13, "A Lighthouse Protocol for Random Microwave Contacts,"* QEX*, Technical Notes, ARRL, July/Aug 2004, p 60 and "How Safe is Your Ham Shack?,"* QST*, ARRL, June 1978, pp 11-13.*

Upcoming Conferences

46th Annual Central States VHF Society Conference

July 25–27, 2013
Elk Grove Village Holiday Inn
1000 Busse Rd (Route 53)
Elk Grove Village, IL, 60007
Reservation Phone: 800-972-2494

The Central States VHF Society, Inc (CSVHFS) invites you to attend the 47th annual conference in Elk Grove Village, IL, July 25–27, 2013. The Planning Committee has a fun-filled, educational event in store for you! The on-line registration form is not yet active, but you can register when you arrive.

The Saturday evening banquet speaker will be ARRL Executive Vice President David Sumner, K1ZZ.

Call for Posters

Conference attendees are invited to bring posters for display. They may be technical or non-technical but will cover the full breadth of amateur weak signal VHF/UHF activities. Topics of Interest include:

- VHF/UHF Antennas, including modeling/design, arrays and control
- Construction of Equipment – such as transmitters, receivers and transverters
- RF power amplifiers – including single and multi-band, vacuum tube and solid state
- Preamplifiers (low noise)
- Regulatory topics
- Software defined radio (SDR)
- Test equipment – including home-brew, using and making measurements
- Operating — including Contesting, Roving and DXpeditions
- Propagation – including ducting, sporadic E, tropospheric and meteor scatter
- Digital Modes – WSJT, JT65 and others
- EME (Moon Bounce).

Non weak signal topics such as FM, repeaters and packet radio are generally not considered, although there are exceptions. If you have any questions about the suitability of a topic, contact K9JK, listed below.

If you would like to display a poster, please contact John Kalenowsky, K9JK, as *soon as possible* with the title and a short description. You can reach John at **csvhfs2013@gmail.com** or 58 N Oak St, Palatine, IL 60067-5238. Author Guidelines and other details are available at the Society website: **www.csvhfs.org**.

The 32nd Annual ARRL and TAPR Digital Communications Conference

September 20-22, 2013
Cedarbrook Lodge
18525 36th Ave South
Seattle, WA 98188-4967
Reservation Phone: 877-515-2176
www.cedarbrooklodge.com

The accommodations at Cedarbrook Lodge for this year's DCC look like they will be outstanding. Make plans now to attend the premier technical conference of the year, the 32nd Annual ARRL and TAPR Digital Communications Conference. When making your reservations, be sure to mention the TAPR Digital Communications Conference for the special negotiated room rate.

Visit the TAPR website under Conferences (**www.tapr.org/dcc.htm**) for more details. Be sure to reserve your room with the hotel and register with TAPR for the Conference.

The ARRL and TAPR Digital Communications Conference is an international forum for radio amateurs to meet, publish their work, and present new ideas and techniques. Presenters and attendees will have the opportunity to exchange ideas and learn about recent hardware and software advances, theories, experimental results, and practical applications.

Topics include, but are not limited to: Software defined radio (SDR), digital voice (D-Star, P25, WinDRM, FDMDV, G4GUO), digital satellite communications, Global Position System (GPS), precision timing, Automatic Packet Reporting System® (APRS), short messaging (a mode of APRS), Digital Signal Processing (DSP), HF digital modes, Internet interoperability with Amateur Radio networks, spread spectrum, IEEE 802.11 and other Part 15 license-exempt systems adaptable for Amateur Radio, using TCP/IP networking over Amateur Radio, mesh and peer to peer wireless networking, emergency and Homeland Defense backup digital communications, using Linux in Amateur Radio, updates on AX.25 and other wireless networking protocols and any topics that advance the Amateur Radio art.

This is a three-day Conference (Friday, Saturday, Sunday). Technical sessions will be presented all day Friday and Saturday. In addition there will be introductory sessions on various topics on Saturday. To get a better idea of DCC content, you can watch the videos from the 2012 DCC on-line at **www.HamRadioNow.tv**.

Join others at the conference for a Friday evening social gathering. A Saturday evening banquet features an invited speaker and concludes with award presentations and prize drawings.

The ever-popular Sunday Seminar has not been finalized yet, but is sure to be an excellent program. This is an in-depth four-hour presentation, where attendees learn from the experts. Check the TAPR website for more information: **www.tapr.org**.

Call for Papers

Technical papers are solicited for presentation and publication in the *Digital Communications Conference Proceedings*. Annual conference proceedings are published by the ARRL. Presentation at the conference is not required for publication. Submission of papers are due by 31 July 2013 and should be submitted to: Maty Weinberg, ARRL, 225 Main Street, Newington, CT 06111, or via the Internet to **maty@arrl.org**. There are full details and specifications about how to format and submit your paper for publication on the TAPR website.

Microwave Update 2013

Morehead, Kentucky.
October 18 – 19, 2013
Space Science Center
Morehead State University
Morehead, Kentucky 95054

Microwave Update (MUD) is an annual event held since 1985. This year the event is hosted at Morehead State University. MUD is a conference dedicated to microwave equipment design, construction, and operation. It is focused on, but not limited to, amateur radio on the microwave bands. There are technical presentations all day Friday and Saturday. There is a pre-conference picnic planned for Thursday October 17, and a Friday evening flea market at the Hampton Inn. A small hamfest is tentatively planned for Sunday October 20.

Few other details were available as this is being written. Check the Microwave Update website for more information: **www.microwaveupdate.org**.

2013 AMSAT Space Symposium and Annual Meeting

Hosted by the Johnson Space Center Amateur Radio Club
November 1 – 3, 2013
Marriott Hobby Airport Hotel
Houston, Texas
Reservation Phone: 713-943-7979

AMSAT announces the 2013 AMSAT Space Symposium will be held on Friday, November 1st through Sunday, November 3rd, 2013.

Call for Papers

Proposals for papers, symposium presentations and poster presentations are invited on any topic of interest to the amateur satellite community. See the AMSAT website for more details: **www.amsat.org**.

Send a tentative title of your presentation as soon as possible, with final copy to be submitted by October 1 for inclusion in the printed proceedings. Abstracts and papers should be sent to Dan Schultz, N8FGV, at **n8fgv@amsat.org**.