Jim Satterwhite, K4HJU

15006 Sawyer Ranch Rd, Austin, TX 78737; **k4hju@arrl.net**

Calibration and Monitoring of Frequency Standards — Phase Method

That frequency standard is only as good as your calibration technique allows. The author describes a monitoring system he uses to keep his under control.

In this article I am speaking primarily to the engineer, technician or experimenter who has an interest in precision frequency standards and their calibration and monitoring. This article will deal with the calibration and monitoring of the various 10 MHz frequency standards. I wanted to measure frequency sources dynamically to an accuracy in the vicinity of one part in 10^{12}. This resulted in designing and building a high performance test set for this purpose. The purpose of the design is to detect a phase difference as quickly as possible to (A) detect short-term drift and (B) minimize measurement time. The performance achieved is an error display noise floor of approximately of 2 parts per trillion.

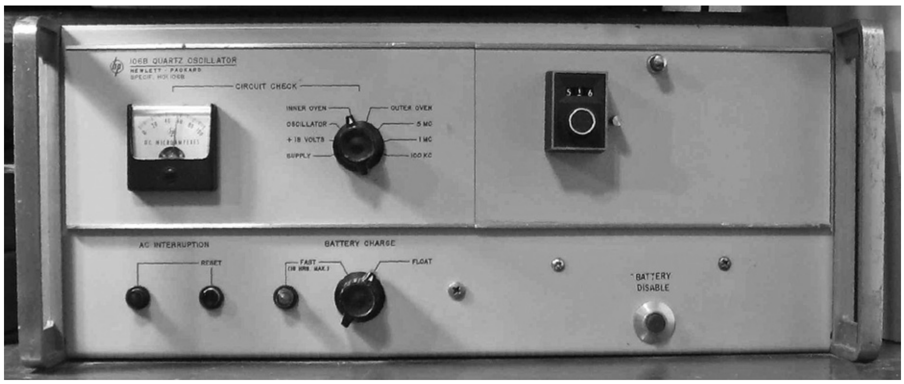


Figure 1 — My HP-106B quartz oscillator.

Years ago, I acquired an HP-106B frequency standard, an early 1960s frequency standard intended as a reference for calibrating other standards at remote locations. It would be calibrated, put on an airplane and taken to the remote location. Then the remote standard would be calibrated using the HP-106B as the reference. The HP-106B contains a 5 MHz crystal oscillator in a double oven. See Figure 1. Frequency standard technology has come a long way since then.

I searched for ways to verify the calibration of this instrument. What is involved and how do I go about doing it? I had designed and built what I consider to be some quite good and versatile frequency counters, and was familiar with frequency measurement; at least to parts in 10^8, but the HP-106B specification is a stability of $\pm 5 \times 10^{-11}$ over 24 hours. When the objective becomes accuracy in the vicinity of one part in 10^{11} or 10^{12}, the challenge increases considerably.

Low-tech ways of calibrating the HP-106B would involve comparing the device under test (DUT) with a standard reference using one of the following methods:

Audio Zero Beat: Listening on a receiver for a zero-beat signal when the output of the device under test is mixed with the signal from an external reference such as WWV.

- Accuracy is poor, because the ear can only detect errors of a few parts per hundred.
- Accuracy is also determined by the low-frequency response of the audio portions of the receiver because phase errors of one or two hertz must be audible to the observer.

Frequency Counter: Directly measuring the frequency of the device under test with a frequency counter.

- Accuracy is limited because most counters have internal references that are accurate to a few parts per million for low end references and tens of parts per billion for high end references.
- Since the HP-106B is capable of precision approaching $\pm 5 \times 10^{-11}$, the reference to which it will be compared must be at least as accurate, and by preference it will be at least an order of magnitude better than this.

Lissajous Pattern: By applying the reference and device under test signals to the X and Y channels of an oscilloscope, a Lissajous pattern can be created. The pattern goes from an ellipse to a 45° line and back to an ellipse as the phase between the reference and device under test varies through 360°. This is quite effective for tuning over a limited range; it is poor for measurement, however.

Table 1
Common Accuracy Terms

Description	*Scientific Notation*
Common Term	*Abbreviation*
One part in 10^6	1×10^{-6}
Parts per million	ppm
One part in 10^9	1×10^{-9}
Parts per billion	ppb
One part in 10^{12}	1×10^{-12}
Parts per trillion	ppt

Table 2
Relative Accuracy of Various Frequency References

Reference	*Modification*	*Accuracy Range*
Crystal	—	1 to 100 ppm
Crystal	Temperature Compensated TCXO	~ 0.1 ppm
Crystal	Ovenized OXCO	0.001 - 0.1 ppm
Crystal	Double Oven	~ 50 ppt
Crystal	GPS supervised OXCO	~ 5 ppt
Rubidium	—	~ 50 ppt
Rubidium	GPS supervised	~ 5 ppt
Cesium	—	0.01 to 0.1 ppt
Hydrogen Maser	Passive	1 ppt
Hydrogen Maser	Active	0.0007 ppt

Reference Signal Sources

In discussing accuracy several terminologies are commonly used. Table 1 summarizes this terminology.

The first part of frequency calibration is to compare the unknown frequency to a frequency known to an accuracy and precision greater than the desired calibration level. That means at the outset you have to have access to a source more accurate than the desired calibration level.

References for frequency measurement cover the gambit from stand-alone crystal oscillators on the low end to Cesium standards and beyond on the high end. Somewhat of an overview is given below for the various reference technologies in Table 2. For the budget concerned experimenter, there are two good choices.

1) Rubidium standards are appearing as surplus items because they are being retired from service in cellular telephone systems around the world. Frequency-stable outputs can be accurate at the 10s of parts per trillion (ppt) level.

2) Crystal oscillators can be "disciplined" — phase-locked — to US Global Positioning System (GPS) satellites.

- The GPS system requires that individual satellite clocks be maintained at accuracies exceeding a few parts per trillion.
- Earth-bound oscillators can be synchronized with the GPS satellite clocks, using timing information derived from the GPS signal.

Calibration

In order to calibrate a standard or device under test, it must first be measured relative to a known reference. Then the task is to adjust the standard or device under test into agreement with the reference.

Measurement

Given the availability of such references, the *metrologist* performing a verification or calibration must answer two important questions:

1) Is a single, brief frequency comparison sufficient to ensure an accurate measurement?

2) In making the measurement, how does one compare two frequency sources and estimate the difference/error between them?

The answer to the first question is No.

- Separate measurements an hour, a day, or a week apart do not confirm that the difference between a reference and a device under test is consistent.
- At best, they can only confirm that the measured error was the same at two points in time. They could not show that the device under test frequency remained constant through the measurement period.
- Many measurements at short intervals over a long time are needed to establish the variation of the device under test with time.
    - The results can be plotted to show, graphically, how the reference and device under test frequencies differed over time.
    - The results can be "averaged" to arrive at a crude figure of merit for the device under test.

A number of answers can be given to the second question:

- A frequency counter with a sufficiently accurate time base can be programmed to accumulate measurements over a period of tens, hundreds, or even thousands of sec-

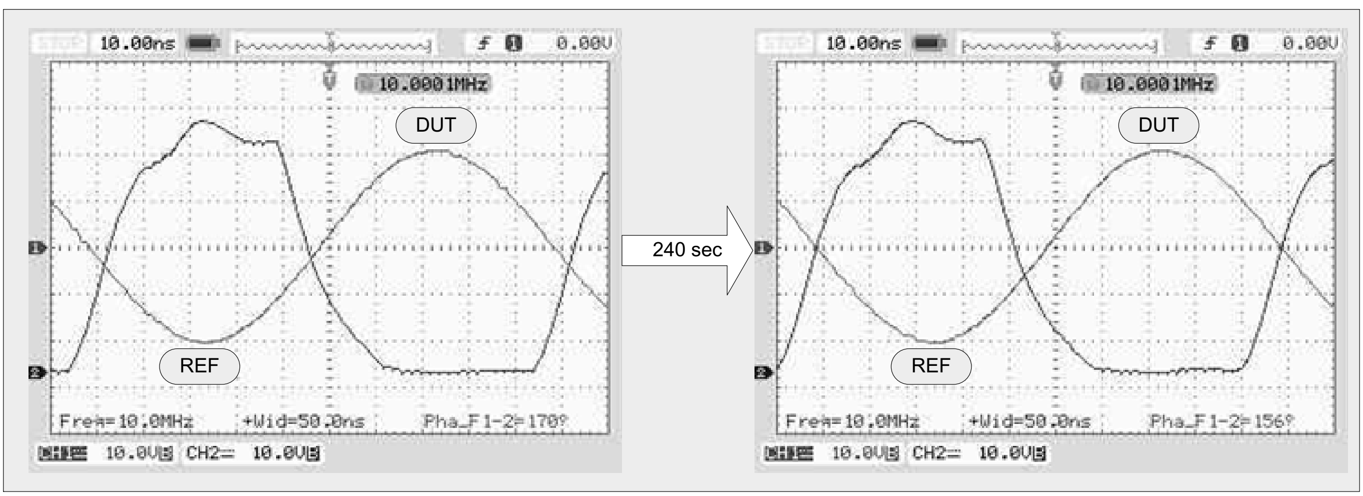


Figure 2 — These screen shots show an example of relative phase measurement with an oscilloscope.

onds. The long measurement time base improves overall accuracy by integrating errors over time; transient deviations are averaged, however.

- Relative phase method: when the frequencies of the reference and the device under test are nearly the same, measuring the difference in the phases of the two signals versus time may also be possible.
  - Oscilloscope — With both reference and device under test signals applied to the inputs of an oscilloscope, the scope can be triggered by the reference signal and the relative phase can be compared at a time A and at a time B. Then with the difference in time between A and B and the difference in the phase readings, either in degrees or in time, the difference in frequency can be calculated. An example is shown in Figure 2. Here the time is 240 seconds and the phase is 156° to 170°, or a shift of 14°, to yield a difference in frequency of 16 ppt. This example is rather straightforward in that the oscilloscope being used here computes phase for us. An eyeball estimation of the phase is somewhat tedious. Also, take note that variations in the phase difference during that time are practically impossible to observe.
  - Strip Chart — My first attempt at verifying the accuracy of my HP-106B standard involved comparing its output to WWVB at 60 kHz and recording the relative phase of the WWVB reference and the device under test on a strip chart recorder.
    - This method can be time consuming because resolving a difference of only 10° at an accuracy of 1×10^{-11} would require 12.9 hours. Increasing the resolution accuracy to 1×10^{-12} would require extending the measurement time to 5.33 days!
    - The measurement time can be reduced when comparisons are done at higher frequencies.

Adjustment

Having measured the frequency difference, the next task in calibration is to adjust the device under test to some given limit, or as close as possible to the reference frequency. When long measurement times are encountered, the adjustment process becomes problematic since the operator cannot see the effect of the adjustment until another measurement is made. It takes time and patience to make this process converge. If the operator can see the effect in real time, the process becomes much easier and straightforward.

Automated Relative Phase Method

Automating the Relative Phase Method for comparing the frequencies of two sources can be extremely accurate if an accurate reference is available.

1) The method determines the rate of change of phase between two sources. A computer monitors the phase difference between them and presents the error graphically.

2) For example, a one-hertz phase error (one 360° phase change per second or 1 Hz frequency difference) represents a phase change of 360° per second. At 10 MHz, the time required to complete one hertz (one RF cycle) is 100 ns, so the error due to phase change can be described as 100 ns per second.

3) We can elaborate the idea by stating the change in degrees of phase error per second for a given fraction of a cycle. Thus for a 10 MHz comparison, Table 3 shows some useful relations.

4) Clearly, small phase errors approaching 0.0036 degrees/second or 0.001 nanosecond/second become visible only after many seconds if they are to be observed on an oscilloscope. A 1 ns error, for example, completes a full 360° phase shift in 1×10^{9} seconds, which is nearly 280 hours, or 11½ days.

4) An important human factor involved with these types of measurements on an oscilloscope is keeping track of the variations over long periods of time.

From Table 3, some useful conversion constants for calculations emerge.

- degrees / s = (1/0.0036) ppt
- degrees /s = (1/360) ppm
- ns / s = 1000 ppt
- ns /s = 0.001 ppm

This method must be automated if accuracies approaching 1 ppt are to be realized.

Monitoring

Something I have found interesting in working with precision oscillators is that rarely are they what they seem at first glance. Since I have designed and built the instrument described in this article, I have found the true performance of several standards I have in my lab to be less ideal than I had expected. Because you calibrate an oscillator to 1.1×10^{-10} and you come back tomorrow and it reads 1.2×10^{-10} doesn't necessarily mean that you can depend on it having 2×10^{-10} accuracy for general measurements. Now that I have the capability of monitoring various sources at the 10^{-11} or 10^{-12} level, I have greater reservations about the performance I can expect from a given source with a certain calibration level. If you want to know how a source will truly perform for a given application, I believe it is essential to monitor the error performance over a period of time, and from that determine what you can count on for the performance of that source.

The Test Set

I chose to accurately measure the relative error between a reference standard of known accuracy and an unknown device under test (DUT), dynamically, and in real time. I determined that the best candidate for this

Phase Comparison at 60 kHz

At 60 kHz, a frequency error of 10 ppt represents a phase change difference between the reference and the device under test of 0.0002156° / s. Thus, to accumulate a phase change difference of 10° would require 46382 seconds or 12.9 hours. Achieving a frequency error of 1 ppt would require 463820 seconds or 5.33 days.

For 10 MHz the times are much shorter. For the same error levels as for 60 kHz, the times would be 278 s (10 ppt) and 2783 s (1ppt).

Of course, resolving the frequency difference change accurately depends on being able to actually see the phase difference change on the medium presented, such as the strip chart or oscilloscope, accurately. That is another matter entirely.

Table 3
10 MHz Error Values Versus Frequency Error

Error (Hz)	*ppm / ppt*	*Degrees /s*	*ns / s*
1	0.1 ppm	360	100
0.1	0.01 ppm	36	10
0.01	0.001 ppm	3.6	1
0.001	100 ppt	0.36	0.1
0.0001	10 ppt	0.036	0.01
0.00001	1 ppt	0.0036	0.001

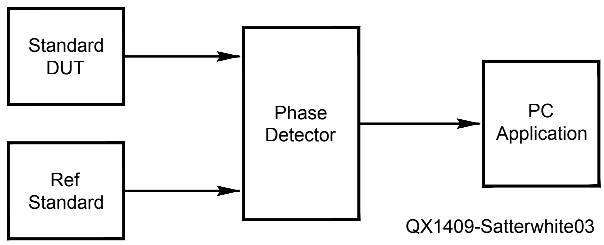


Figure 3 — This block diagram illustrates the basic concept of the phase method of calibrating the oscillator.

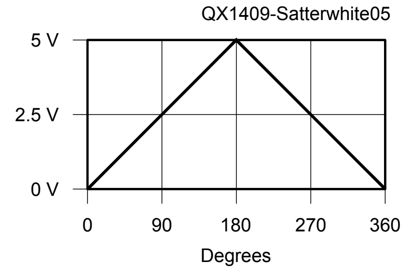


Figure 5 — This graph shows a single XOR gate phase detector output.

measurement was an accurate phase measurement and automation of the method with as much precision as I could bring to bear on the implementation. Further, I wanted to present the results in real time on a computer screen in strip chart form for observation. In this way I could observe the real time performance of various standards. What follows is the result of the evolution of that development. This design limits the input frequencies to 10 MHz for both the reference and the device under test with one exception. The device under test input can accept a 5 MHz input by moving a jumper.

Overview

As shown in Figure 3, the basic concept is simple. The specifics get a little more interesting.

The figure shows that the phase of the reference and device under test are compared and a value relating to the phase difference is passed to the computer. The computer software initiates a fixed time interval trigger for each phase measurement. Calculations that result in the computed error being displayed on the computer screen are made from the fixed time interval and the measured phase difference for each interval.

Figure 4 shows the detailed block diagram of the test set. To achieve high accuracy, the signals must be conditioned significantly. Each channel is first detected and phase-locked to a higher frequency crystal VCO to obtain a precise square wave for presentation to the phase detector. This is necessary because, in simply squaring the input signals, it is difficult to maintain the necessary 50% duty cycle required for precise phase determination. In addition, for the quadrature phase detector chosen, the reference frequency needs to be four times the reference frequency. This will be discussed below.

The reference and device under test channels are almost identical. The exceptions are that the reference channel operates at 40 MHz and its output supplies clocks for the quadrature phase detector. I will discuss the quadrature phase detector below.

The reference input signal is converted from analog to digital form by the input

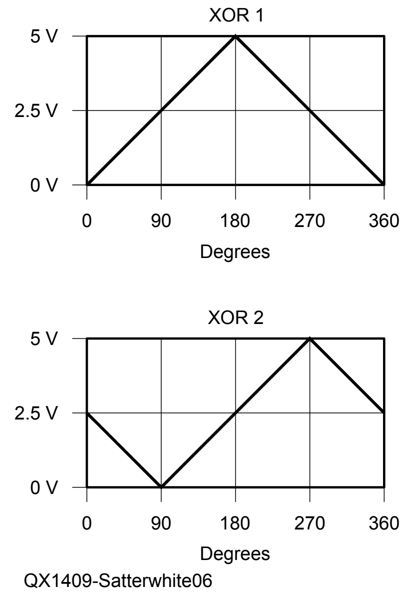


Figure 6 — Here is a graph of the Quadrature XOR phase detector outputs.

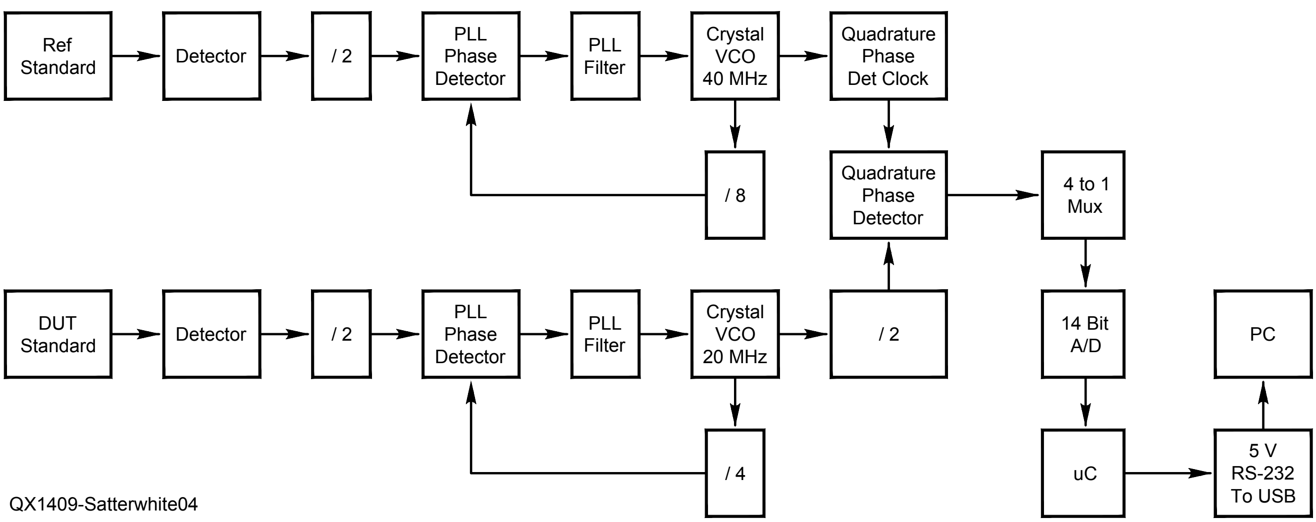


Figure 4 — Here is a detailed block diagram of the system.

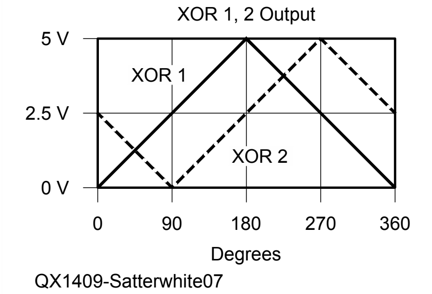


Figure 7 — This graph is an overlay of the Quadrature XOR phase detector outputs.

detector and is then divided by two by a digital counter. The 5 MHz output feeds the 74HC4046 PLL phase detector. I prefer to operate this PLL phase detector at somewhat less than 10 MHz. The output of the PLL phase detector feeds the crystal VCO via the PLL filter. The 40 MHz VCO output feeds a divide by eight circuit, which supplies a 5 MHz square wave to the PLL phase detector to close the 40 MHz PLL loop. It also drives a circuit to derive the reference clocks for the quadrature phase detector.

The device under test channel is identical to the reference channel, with two exceptions. The crystal VCO is operated at 20 MHz and the output is divided by two to feed the quadrature phase detector at 10 MHz and divided by four to feed the PLL phase detector at 5 MHz, closing the 5 MHz PLL loop. The 74HC4046 PLL phase detector is fed from the input divided by two for a 10 MHz device under test input or moving a jumper bypass the divide by two for a 5 MHz device under test input.

The outputs of the reference and device under test channels are compared in the quadrature phase detector (QPD) and fed to a 14 bit A/D that is coupled to the microcontroller via the Serial Peripheral Interface (SPI). The microcontroller takes in the digitized phase values and routes them to the computer via a 5 V RS-232C-to-USB adapter. The computer, running a *Windows* application written in National Instruments *Lab Windows/CVI*, processes the data and presents the analyzed results on the computer monitor.

Quadrature Phase Detector

The phase detector chosen for this implementation is the Exclusive-OR (XOR) logic gate. For a single XOR, the output waveform for a phase difference from 0° to 360° is shown in Figure 5. The output, when filtered, is a triangular wave whose amplitude varies with the phase difference between the two inputs. Note that the output is multi-valued, that is there are two phase values for each voltage value, therefore the unique phase cannot be determined from the voltage reading without additional information.

This can be remedied by the addition of a second XOR gate driven by the reference signal, which is 90° out of phase and which forms the quadrature phase detector. The quadrature phase detector reference signals are derived from the input reference clock multiplied by four. The result is shown in Figure 6 and overlaid on the same trace in Figure 7.

From the combination of the two outputs, the in-phase component (I) and the quadrature component (Q), the unique phase can be determined, since there is a unique value pair for each phase point.

The circuit I use for the generation of the quadrature clocks is shown in Figure 13. It uses two additional XORs to help remove a very slight nonlinearity of phase versus amplitude at the peaks and valleys of the output. For this, I generate 10 MHz clocks at 0° (I), 90° (Q), 180° (–I) and 270° (–Q). Figure 8 shows the XOR outputs using the four clocks. I and Q are the in-phase and quadrature phase components.

The I and –I and the Q and –Q components are combined in software to give the resultant I and Q from which the phase calculations are made. From the I and Q, the phase value is determined. At a sample rate of 10 samples/second and given the calculated phase values, the phase change rate is determined. The phase change rate multiplied by a conversion factor is the basis for the short-term error. The long-term error is

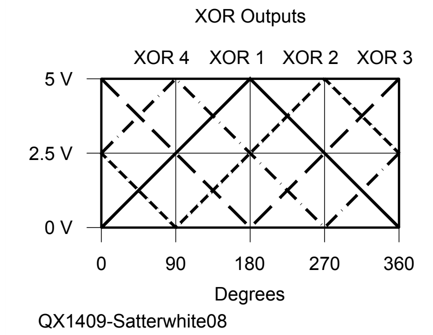


Figure 8 — Outputs of the four XORs versus device under test phase.

GUI Trace Description

The GUI is in color and the Figures are presented without color. On the GUI, the red/pink trace is the short-term error and the blue/light blue trace is the total accumulated averaged error since the test began. Since negative values are problematic on a log chart, the error values are differentiated for sign by color shading, with the dark colors representing positive error and the light shading being the negative error. The green trace is the accumulated phase error in degrees. For this display the phase display is limited to values between –360° and +360°, so that we see the trend in phase with time.

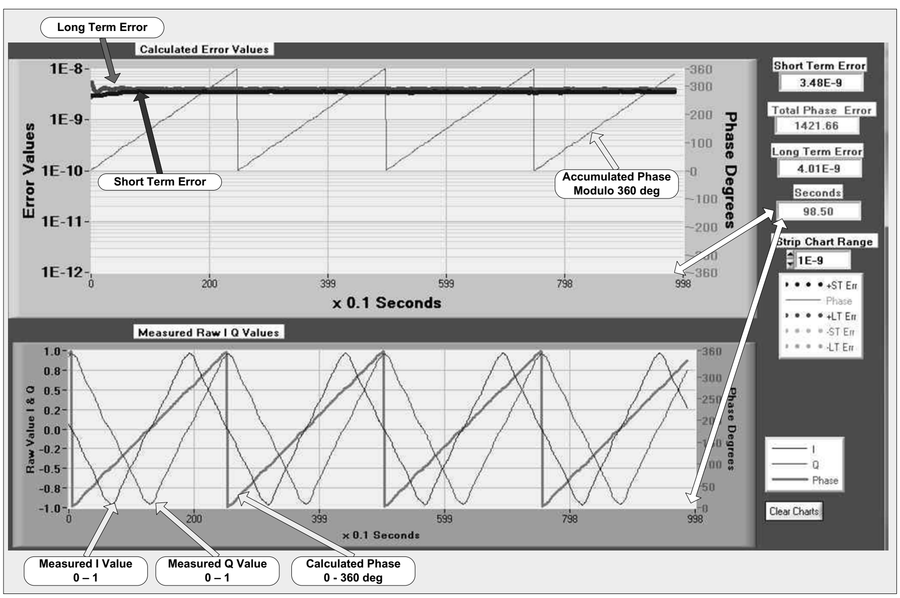


Figure 9 — Here is an example of the Quadrature phase detector output.

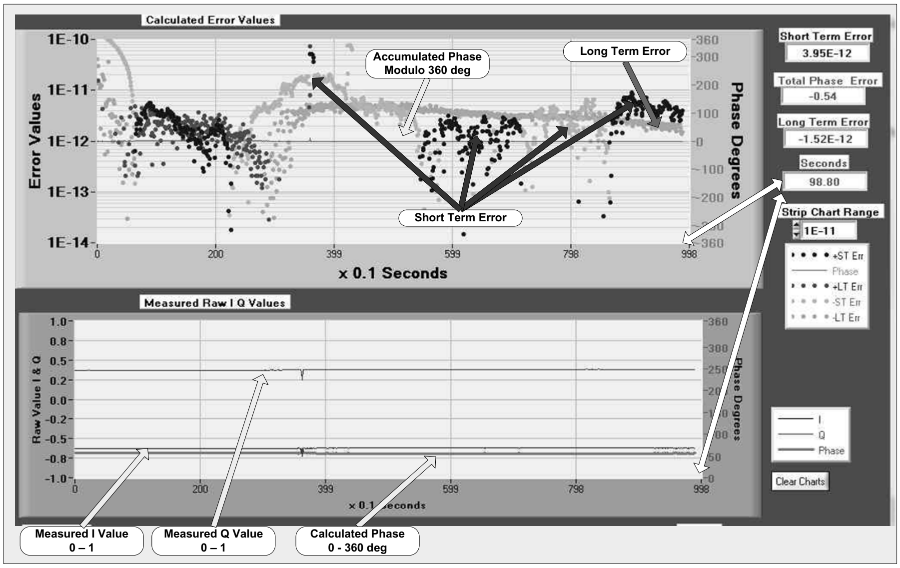


Figure 10 — This is an example of the output data.

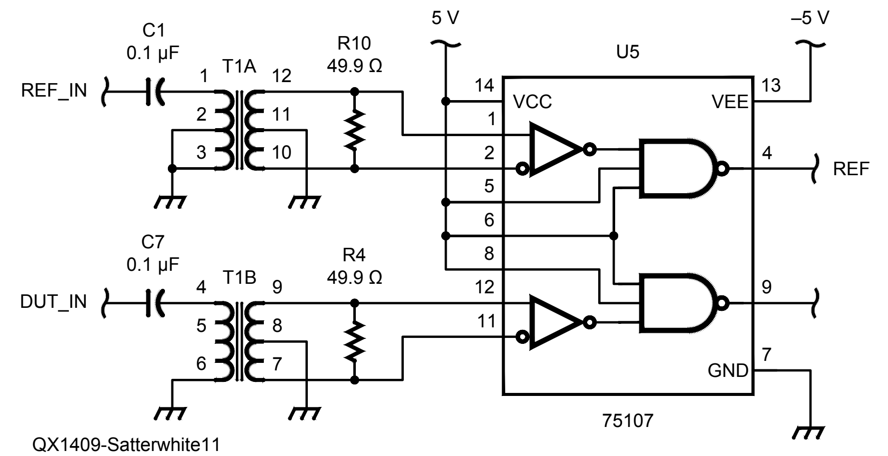


Figure 11 — This is a schematic diagram of the input detector.

calculated by taking the first phase reading in a given run as the start reference. Then the present phase difference minus the start reference divided by the total time since start is the long-term degrees/second that times the conversion factor is the long term error.

Figure 9 shows a screen presentation from the GUI for a period of 98.5 s. A Trimble Thunderbolt GPS-supervised source is used as the reference input and an ovenized crystal oscillator as the device under test input to the test set. There are two strip chart presentations. The lower strip chart shows the measured I and Q values changing with time and the phase calculated from the two. From the calculated phase, the short-term (incremental phase change versus time increment) and the long-term (total phase change versus total time) errors along with the accumulated phase are calculated and presented in the upper strip chart in the figure. In the GUI, colors are used to identify the various traces. Here, the long-term error is 4.01 ppb over a period of 98.5 s, and the last increment of short-term error is 3.48 ppb.

Figure 10 shows an annotated screen presentation of measurements using a Trimble Thunderbolt GPS-supervised source as the reference and a rubidium oscillator as the device under test. Here the phase movement is so slow that the quadrature phase detector outputs look almost stationary (lower strip chart). The top strip chart shows the short-term and long-term error traces and the accumulated phase difference trace.

Note that the scale on the left of the upper strip chart has been adjusted to better show the excursions of the traces. If we study the traces closely we can determine the extent of the long-term error trace as it stabilizes and remains rather convergent while the short-term trace varies considerably. This test is approaching the limit of the test set, as can be seen below in the section on performance. Note that the total phase error for the 98.8 s is only –0.54°, giving a long-term error of –1.52 ppt.

Processing

High Level Circuit Description

The input detector (Figure 11) is designed to accept a variety of signal input levels and types. The transformer provides ground isolation between the two sources. An attempt has been made to isolate the reference and device under test inputs as much as possible to reduce the possibility of entraining of either of the two sources. Some oscillators exhibit the possibility of entraining with minute cross-coupling from a source close to their natural frequency and these two sources are quite close in frequency. The 75107 line receiver provides a reliable transition detector function.

The VCO does not necessarily give a symmetric waveform output, so a divide by two is necessary to guarantee essential symmetry. Symmetry is particularly important in driving the phase detectors to derive phase as precisely as possible.

The quadrature phase detector shown in Figure 12 offers the ability to uniquely identify the phase difference from 0° to 360°. Only two XORs are necessary to accomplish this; however the second two allow a bit more resolution, as will be discussed below.

Figure 12 shows the quadrature phase detector. The device under test signal is compared to the reference signal at 0°, 90°, 180°, and 270° to give the Q, I, –Q and –I signals respectively. Each XOR detector is followed by a pull-down resistor to –5 V and an RC low pass filter. The detector outputs go through an analog multiplexer to the A/D converter. The pull-down resistor to –5 V helps increase the sharpness of the minimum value of the detector output.

To ensure symmetry, the circuit must be driven by four times the output frequency. Figure 13 shows the circuit for the generation of the quadrature phase detector clocks. Starting with the clock output of the 40 MHz crystal VCO that is phase locked to the 10 MHz reference input, the 40 MHz clock drives the four stage synchronous counter to derive the 5 MHz PLL feedback and the D flip-flops to derive the four quadrature phase detector reference clocks.

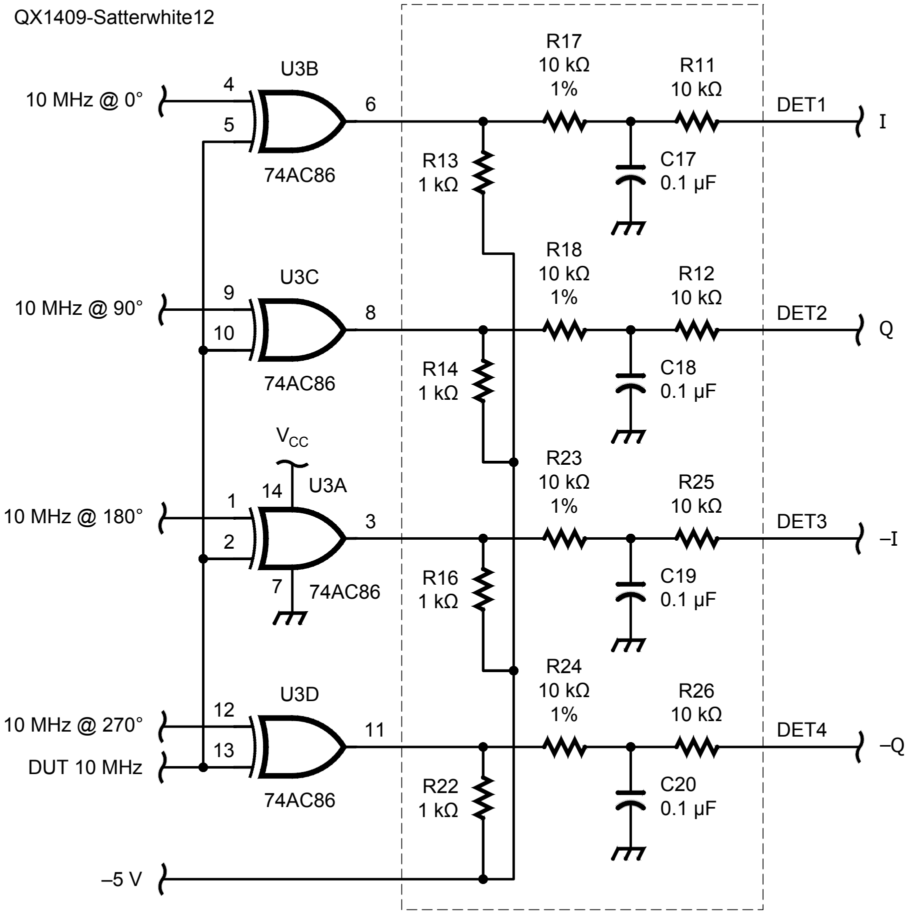


Figure 12 — Here is the schematic of the Quadrature phase detector.

Practical Considerations Limiting Performance

A number of practical considerations limit the performance of the over all system. It is interesting to look at the parameters that limit the performance of the test set. Of course the primary limit is the accuracy and stability of the reference source. Beyond that is the accuracy and stability of the PLLs. Further the short-term accuracy is particularly dependent on the resolution of the A/D.

Reference Signal Sources

It is important to note that for a meaningful reading, the reference accuracy must be at least an order of magnitude better than the desired level of measurement. High-accuracy reference sources are available, but most are expensive and for the average experimenter, cost is important. The Trimble Thunderbolt GPS supervised source and various Rubidium sources are available from surplus sources.

Trimble Thunderbolt

There are at least two kinds of GPS-supervised sources. The one commonly seen in experimenter circles uses the GPS one-pulse-per-second (1PPS) output, that is compared with an ovenized crystal oscillator or a Rubidium source and used to derive a signal to correct the source. The second, which is used by Trimble in the Thunderbolt derives the GPS receiver reference clock from an ovenized crystal oscillator. The timing error from the GPS position solution corrects the

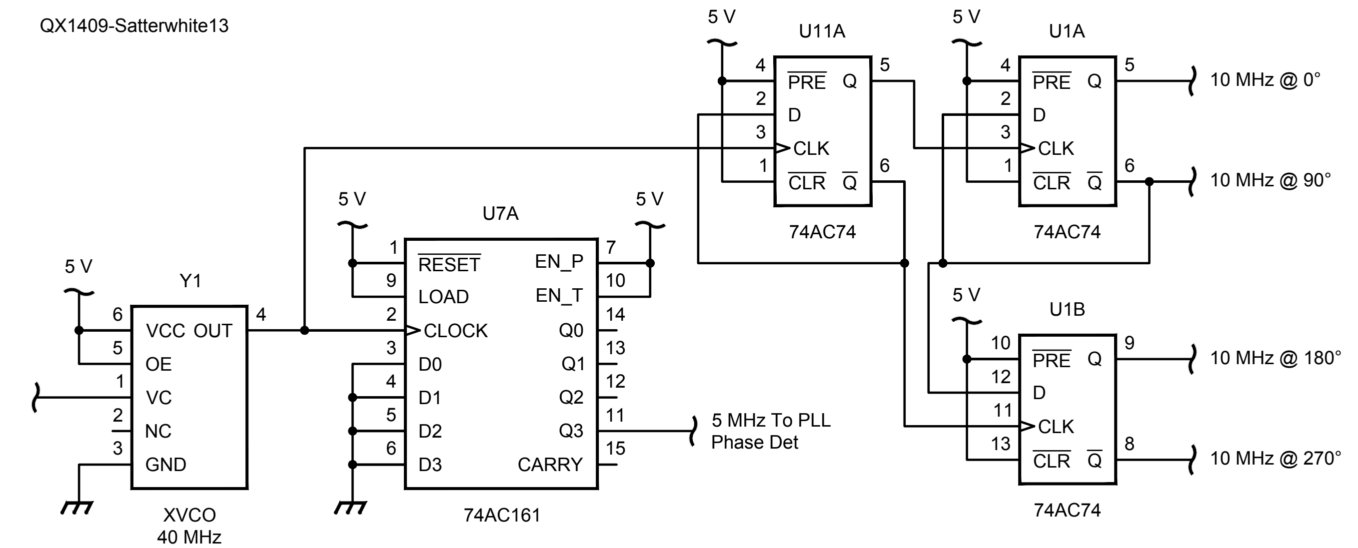


Figure 13 — This is the Quadrature clock circuit.

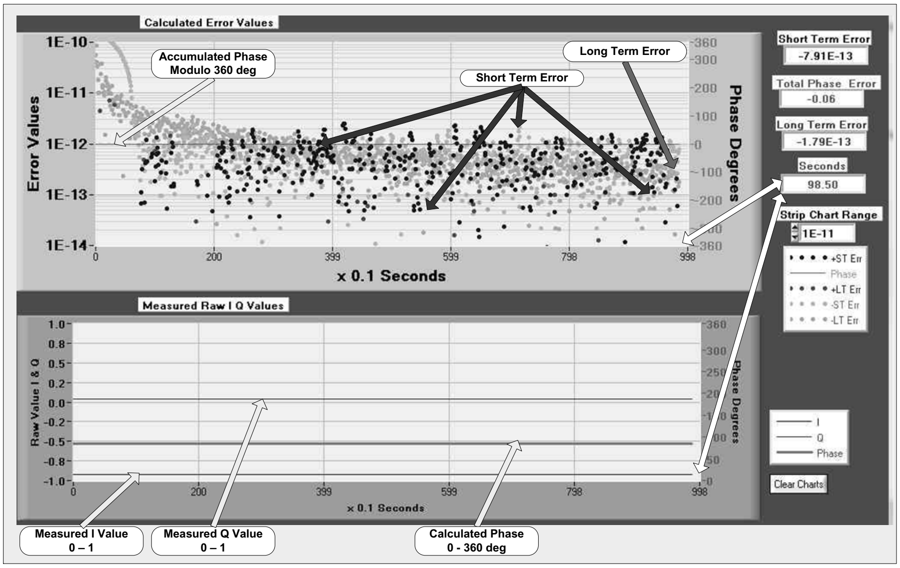


Figure 14 — Here is the data from a Trimble Thunderbolt to Reference and device under test Inputs.

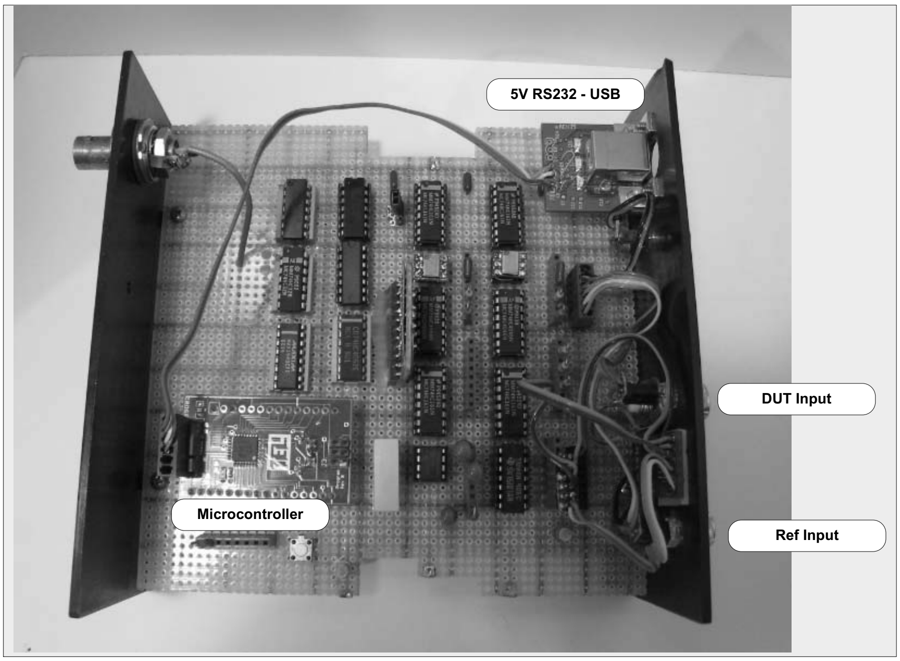


Figure 15 — This photo is a component view of the test set.

ovenized crystal oscillator. The problem with the 1 pulse per second technique is that it is difficult to determine the difference between the supervised source and the 1 pps signal accurately, so making accurate corrections is problematic.

It is interesting that Trimble does not specify the accuracy of the Thunderbolt 10 MHz output, but they do specify the phase noise. I suspect the lack of an accuracy statement is due to the various potential user-related situations where a solid GPS solution is not possible or is intermittent. Such situations might include the setup or the antenna placement. The *Windows* utility they supply to set up and monitor the unit in operation does give a dynamic error report and with the unit I have that error varies around parts in 10^{-12}. Some thought about the stability of these types of sources reveals that at some level they must jitter around. Consider that the source is an ovenized crystal oscillator (VCXO) and with a very few exceptions it is very difficult to make them perform better than 1×10^{-10}. (The exceptions mentioned refer to ovenized crystal oscillators built as instruments such as the HP-106B that can perform to a higher level.) To me that means that in this application, with the VCXO controlled to $X \times 10^{-12}$, it is constantly moving away from that point and must be constantly brought back by the controller. This is variation in frequency about the desired frequency.

This approach works, although I can't verify it because the Thunderbolt is the most accurate source that I possess. For my work, I consider the Thunderbolt to be able to hold something like $< 5 \times 10^{-12}$.

Phase-Locked Loops

The PLLs must very closely follow the reference and device under test sources for an accurate short-term measurement. In a PLL, the VCO output is constantly compared to the reference frequency and an error correction signal constantly controls the VCO. The error correction signal will have some noise and there will also be some error. If the PLL is locked, this only means the VCO phase stays in a given region relative to the reference. There is always some jitter or variation. The objective here is to keep that jitter very, very small. As discussed below in the Performance section, the floor of the test set is approximately less than 2×10^{-12}.

Long-term measurements are less critical because, although the PLL VCO may jitter about the input frequency, it is still locked and the longer-term phase difference is still valid even though the short-term phase difference may have significant error.

A/D Converter

The resolution of the A/D converter is very important. The A/D employed resolves 14 bits, thus allowing the least significant bit to be one part in 16384. Since the range of one slope of the quadrature phase detector output is 90°, one least significant bit represents 0.0055° of resolution. From Table 3, we see that 1 ppt results in a phase difference of 0.0036° / s. Thus, this A/D cannot resolve a one second error of 1 ppt. To improve this resolution, a number of readings are taken for each point and averaged, to yield the value used in the calculations. Resolution could be improved by employing a 16-bit A/D. Resolution is much more important than absolute accuracy, because the calculations are based on difference measurements.

Performance

As is sometimes said, the proof of the

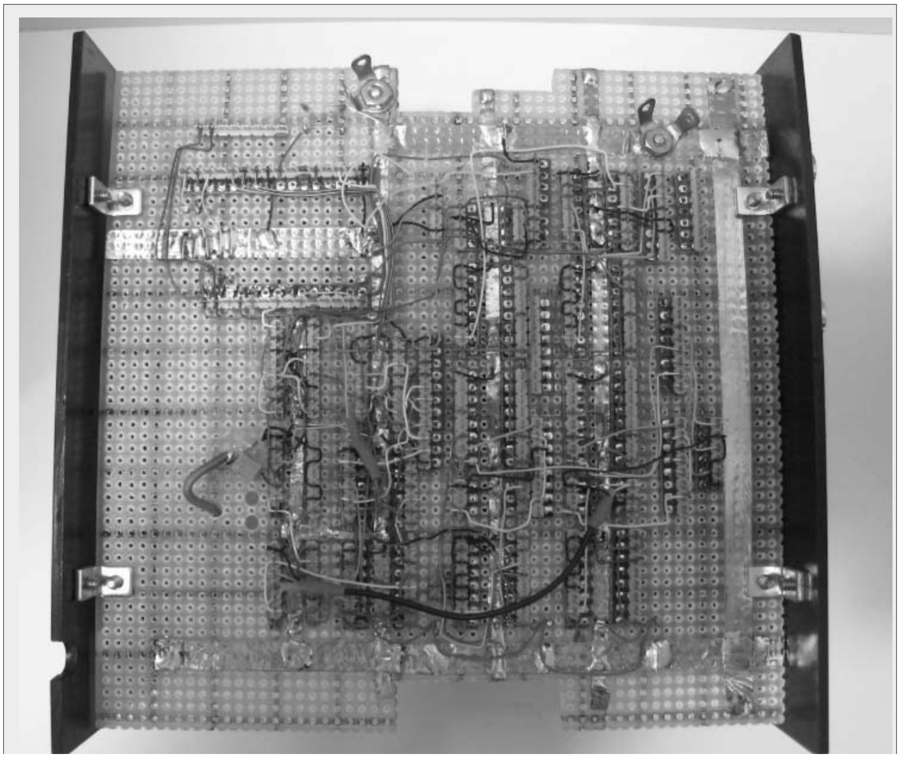


Figure 16 —Test set wiring view.

Figure 17 —Test set front panel view.

pudding is in the eating. All of the factors limiting the range of the test set come into play when both the reference and device under test inputs are connected to the same high-performance source and establish the noise floor of the test set. Figure 14 shows the result of feeding the test set with the same source on both the reference and device under test inputs. This means the two PLLs are operating at the same frequency independently and that the control dynamics of each crystal VCO PLL loop are independent. For the result shown, the two PLLs have to be operating independently with accuracies on the order of 1×10^{-12}. Otherwise, the phase difference between the two channels would be much greater. One factor that does not come into consideration here is that the two are operating from the same frequency and therefore are not moving relative to each other. I have observed in earlier layouts, some interaction between the two channels when the zero crossings of the two channels are in the same region in time. Notice in this case, that the long-term error is predominately positive and that it begins to settle out as time progresses, while the short-term error continues to vary. Again, I beg your indulgence in interpreting the presented data as it is difficult to discern the traces from the black and white image.

From this, it can be seen that the noise floor of the test set is approximately 2×10^{-12}.

Hardware and Software

The hardware is constructed using breadboard techniques and thru-hole DIP ICs, as shown in Figures 15, 16, and 17. The interconnections are made with insulation displacement wiring. The microcontroller is an Atmel ATMEGA328P in a custom module.

The microcontroller code is written in *C* for the Atmel ATMEGA 328P and the *Windows* application is written in *C* using National Instruments *Lab Windows/CVI*.

I created the schematic using *Eagle* software. Figures 18 and 19 show the schematic diagram.

Conclusions

The Test Set has met or exceeded my design goals. This project has given me a much deeper understanding of the performance and also the "care and feeding" of high accuracy frequency reference sources. I have found it interesting to discover just how these high accuracy sources vary with time. Visibility into the operation in real or near real time of a high accuracy standard has given me insight that allows more understanding. One point that has emerged related to the control circuits that allow external adjustment of the sources is that the control circuits need to be of high precision and stability as well.

I have included links to several technical papers about frequency source stability and analysis in the References section.

Questions remain, and in getting here ideas have come forth to improve the test and to include more analysis such as Allan Variance.

Acknowledgement

I would like to offer a special note of thanks to Ken Burch and especially to Brad Rehm, KV5V, for their in-depth reviews, patience and suggestions.

Jim Satterwhite, K4HJU, an ARRL Member, was first licensed in 1956 as KN4HJU. He holds an FCC General Radiotelephone Operator's License, and is a registered Professional Engineer in the state of North Carolina. He is an iNarte Certified EMC Engineer and is a Life Senior Member of IEEE. He received a BEE degree from the University of Florida in 1965 and an MSEE degree from Purdue University in 1966. Jim has been in electronics development and research for the better part of 60 years, including 12 years as a member of the technical staff at Bell Labs

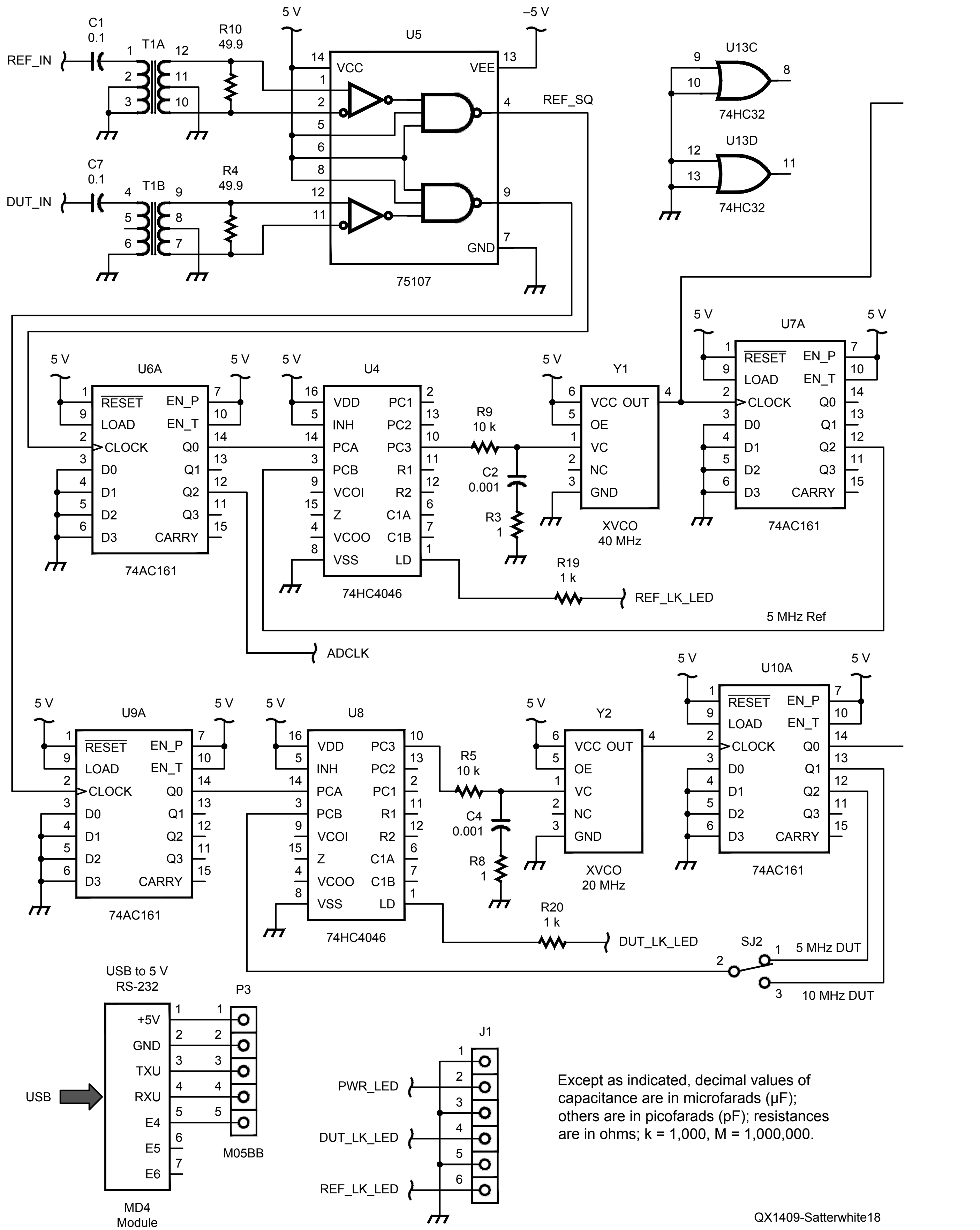
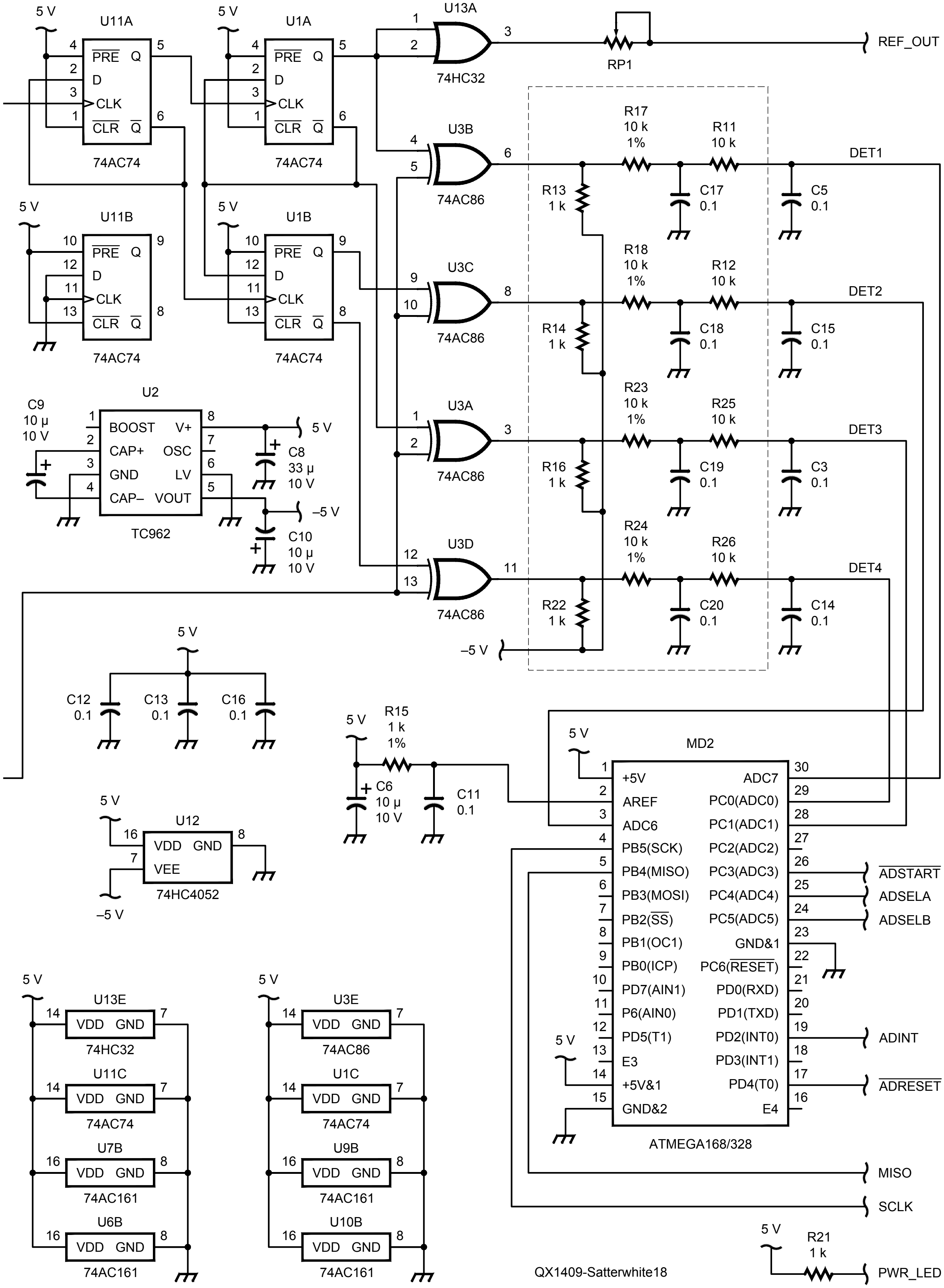


Figure 18 — Here is the schematic diagram of the test set, Part A.


5 V
U11A
PRE Q
D
CLK
CLR Q
74AC74
U1A
PRE Q
D
CLK
CLR Q
74AC74
U13A
74HC32
RP1
REF_OUT
U3B
74AC86
R17
10 k
1%
R11
10 k
DET1
R13
1 k
C17
0.1
C5
0.1
U11B
PRE Q
D
CLK
CLR Q
74AC74
U1B
PRE Q
D
CLK
CLR Q
74AC74
U3C
74AC86
R18
10 k
1%
R12
10 k
DET2
R14
1 k
C18
0.1
C15
0.1
U2
C9
10 µ
10 V
BOOST V+
CAP+ OSC
GND LV
CAP– VOUT
TC962
C8
33 µ
10 V
–5 V
C10
10 µ
10 V
U3A
74AC86
R23
10 k
1%
R25
10 k
DET3
R16
1 k
C19
0.1
C3
0.1
U3D
74AC86
R24
10 k
1%
R26
10 k
DET4
R22
1 k
C20
0.1
C14
0.1
–5 V
C12
0.1
C13
0.1
C16
0.1
R15
1 k
1%
C6
10 µ
10 V
C11
0.1
MD2
+5V ADC7
AREF PC0(ADC0)
ADC6 PC1(ADC1)
PB5(SCK) PC2(ADC2)
PB4(MISO) PC3(ADC3)
PB3(MOSI) PC4(ADC4)
PB2(SS) PC5(ADC5)
PB1(OC1) GND&1
PB0(ICP) PC6(RESET)
PD7(AIN1) PD0(RXD)
P6(AIN0) PD1(TXD)
PD5(T1) PD2(INT0)
E3 PD3(INT1)
+5V&1 PD4(T0)
GND&2 E4
ATMEGA168/328
ADSTART
ADSELA
ADSELB
ADINT
ADRESET
MISO
SCLK
U12
VDD GND
VEE
74HC4052
–5 V
U13E
VDD GND
74HC32
U11C
VDD GND
74AC74
U7B
VDD GND
74AC161
U6B
VDD GND
74AC161
U3E
VDD GND
74AC86
U1C
VDD GND
74AC74
U9B
VDD GND
74AC161
U10B
VDD GND
74AC161
R21
1 k
PWR_LED
QX1409-Satterwhite18

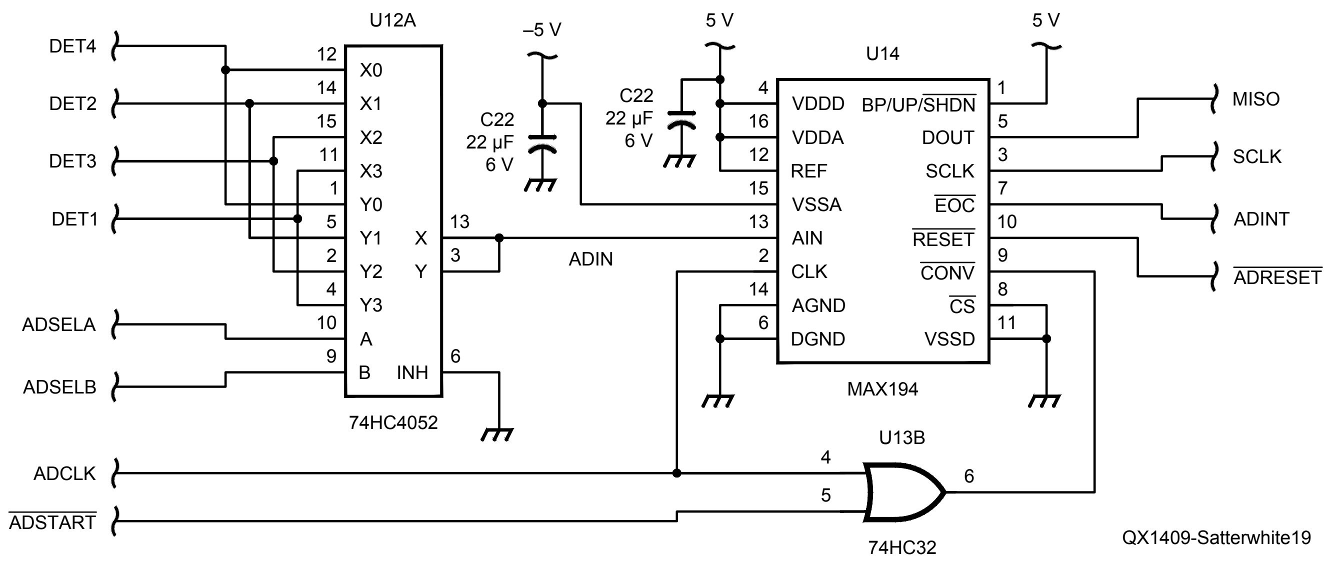


Figure 19 — This is the test set schematic, Part B.

and 32 years as a development engineer with Teltest Electronics, the company he founded in 1982. He holds a number of patents and patent applications. Jim enjoys developing electronic systems and is more comfortable with a track-ball or soldering iron than a microphone.

References

"A Note about Rubidium Oscillators", *Datum*, 12/10/1998, **www.amtestpl.com/DATUM/GmbH/Rubidium-Note-e.pdf**

D. B. Sullivan, D.W. Allan, D.A. Howe, and F. L. Walls, "Characterization of Clocks and Oscillators", *NIST Technical Note 1337*, (March 1990), **http://tf.nist.gov/timefreq/general/pdf/868.pdf**

D. A. Howe, D.W. Allan, and J.A. Barnes, "Properties of Signal Sources and Measurement Methods," *Proceedings of the 35th Symposium on Frequency Control*, **http://tf.nist.gov/timefreq/general/pdf/554.pdf**

W. J. Riley, *Handbook of Frequency Stability Analysis*, NIST Special Publication 1065, July, 2008, **http://tf.nist.gov/timefreq/general/pdf/2220.pdf**

David A. Howe, "Frequency Domain Stability Measurements: A Tutorial Introduction", *NBS Tech Note 679*, March, 1976, **www.adretelectronique.fr/anciens/photos/jacques_rutman/NBS_technical_note_679.pdf**