Technical Description of

5.8 GHz RANGEFINDER

by

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Modifying the LLNL Modular Rangefinder for Operation at 5.8GHz

Introduction

This note describes modifications to a standard LLNL (Lawrence Livermore National Lab) modular 6.4GHz pulse-echo radar rangefinder that was available from LLNL as a technology demonstration unit for licensing purposes. The unit described in this Note was built by and purchased from an LLNL licensee. It is but one of many configurations of a "Swiss Army Knife" modular system designed to demonstrate the flexibility and capabilities of the LLNL patent portfolio. Uses for the modular system include an electronic dipstick, an impulse radar imaging system, fluid level sensing using RF pulse-echo radar, and an electronic trip wire. The specifications for the configuration described herein are

- 2 to 62-foot operational range (0.6 to 18.9-meters)
- better than 0.1% of full scale timing accuracy over temperature
- 1mm range resolution
- FCC Part 15 compliance prescreening at 5.8GHz.

Applications for this configuration include (1) liquid level measurement in storage tanks, (2) river and lake level sensing (environmental sensing), (3) industrial vat and process control, (4) range-gated beam interrupter systems, and countless other applications requiring high resolution radar. A detailed list of the modifications made for this configuration is available only to LLNL MIR licensees.

Brief Description of the Modular Rangefinder

The modular rangefinder is a sampling-type pulse-echo radar operating in the 5.8GHz band. It is comprised of universal modules that are part of a modular system developed at LLNL for demonstration purposes. Its operation can be understood with reference to the block diagram shown immediately below.

A 1.25MHz pulse repetition frequency (PRF) clock drives a delay locked loop (DLL). The DLL splits the clock input into two outputs—a TX and an RX clock. These two clocks have a precise phase relation that is continuously swept by a voltage from the ramp generator. Typically, the TX-RX clock phase, or timing difference, is very linearly swept from 0 to 130ns every 100-milliseconds. Each negative edge of the 1.25MHz TX clock causes the TX module to transmit a short, 2ns-wide RF burst at 5.8GHz. Echo pulses arriving at the receiver module are gated and peak sampled by the receiver module with swept gate timing from the RX clock.

The output of the receiver module is a sampled, expanded time replica of the received echoes. To compensate propagation losses, a variable gain amplifier (VGA) increases the receive path gain as the receive gate is increased in range. The signal from the VGA is rectified and smoothed (envelope detected) and applied to automatic threshold detectors. A digital pulse width modulated (PWM) output is produced by the detectors with a width that increases linearly and very accurately with target range. The PWM scale factor is typically 5-milliseconds per meter, so digitization of range is easily accomplished with a simple counter (not shown).

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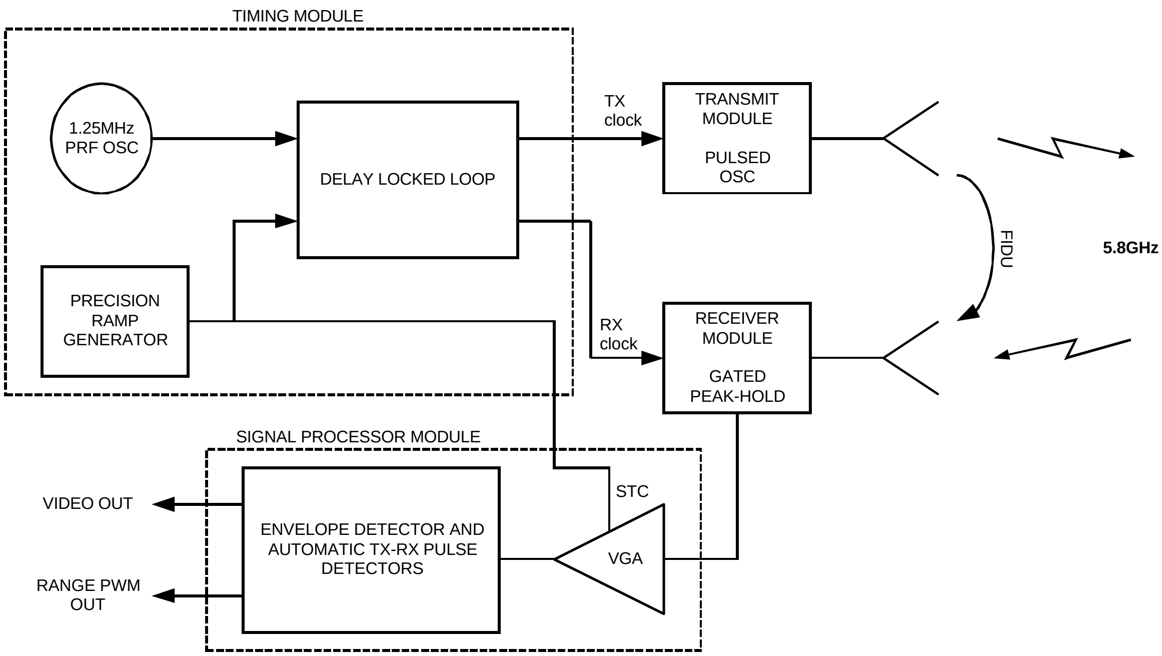


SAMPLING-TYPE PULSE-ECHO RANGEFINDER

The accuracy of the system is defined by timing scale factor errors, timing offset drift, timing nonlinearities such as digital crosstalk, and baseline clutter in the radar return. This note will show that scale factor errors and glitch errors are negligible at the 0.1% level, and that timing offset errors are controlled by the use of a timing reference or fiducial derived from the TX-RX main bang coupling. Errors due to clutter are a function of the radar scene and will not be addressed in this note.

FCC considerations

The rangefinder emits a very short RF pulse having a carrier frequency centered at 5800MHz and a bandwidth that is related to the inverse of the pulse width. While the FCC Part 15 regulations do not place a limit on minimum RF pulse width (there is a prohibition against carrier-free or impulse transmissions) the spectral characteristics of the pulse must comply with Part 15 regulations.

Since the FCC employs a spectrum analyzer of limited bandwidth (~1MHz) compared to the RF emission width (~500MHz), a correction factor must be applied to yield a number indicative of the true RF carrier amplitude, which is what the FCC regulates. This correction factor is called the pulse desensitization correction factor (PDCF),

$$\text{PDCF} = -20\log(1.5BT), \quad \text{where } B = \text{analyzer bandwidth and } T = \text{RF pulse width.}$$

(see HP Application Note 150-2 for details)

The PDCF (in decibels) is added to the observed RF carrier and its harmonics to get the true peak carrier and harmonic amplitudes. Sideband and spurious emissions are not subject to PDCF (according to an FCC communication). Several 5.8GHz pulsed-RF emitters have already been FCC Part 15 approved for LLNL MIR licensees (see FCC the website at www.fcc.gov).

For the system described herein (with moderate gain antennas), its small GaAsFET transmit oscillator is barely capable of exceeding the peak carrier limit in the 5.8GHz ISM band, and it is generally incapable of exceeding permissible sideband levels in the adjacent 15.209 bands (the low power device or LPD bands). The challenge is (1) to reduce PD-corrected harmonic levels at 11.6GHz, 17.4GHz and higher and (2) to ensure there are no intentional emissions in the restricted bands, particularly at 5.4 and 7.2GHz. As will be seen below, a shaped 2ns wide pulse meets this requirement. Faster pulses can be transmitted using high performance RF filters, or with other techniques beyond the scope of this note.

Transmitter Modifications

The LLNL-based demo unit employs an RF transmitter having a GaAsFET pulsed oscillator which is tuned by a discrete inductance in the gate lead. Discrete element tuning is preferred over microstrip resonator tuning due to a reduced tendency for a short pulse oscillator to spuriously oscillate at a harmonic. Discrete element tuning can also be designed to be independent of PCB parameters.

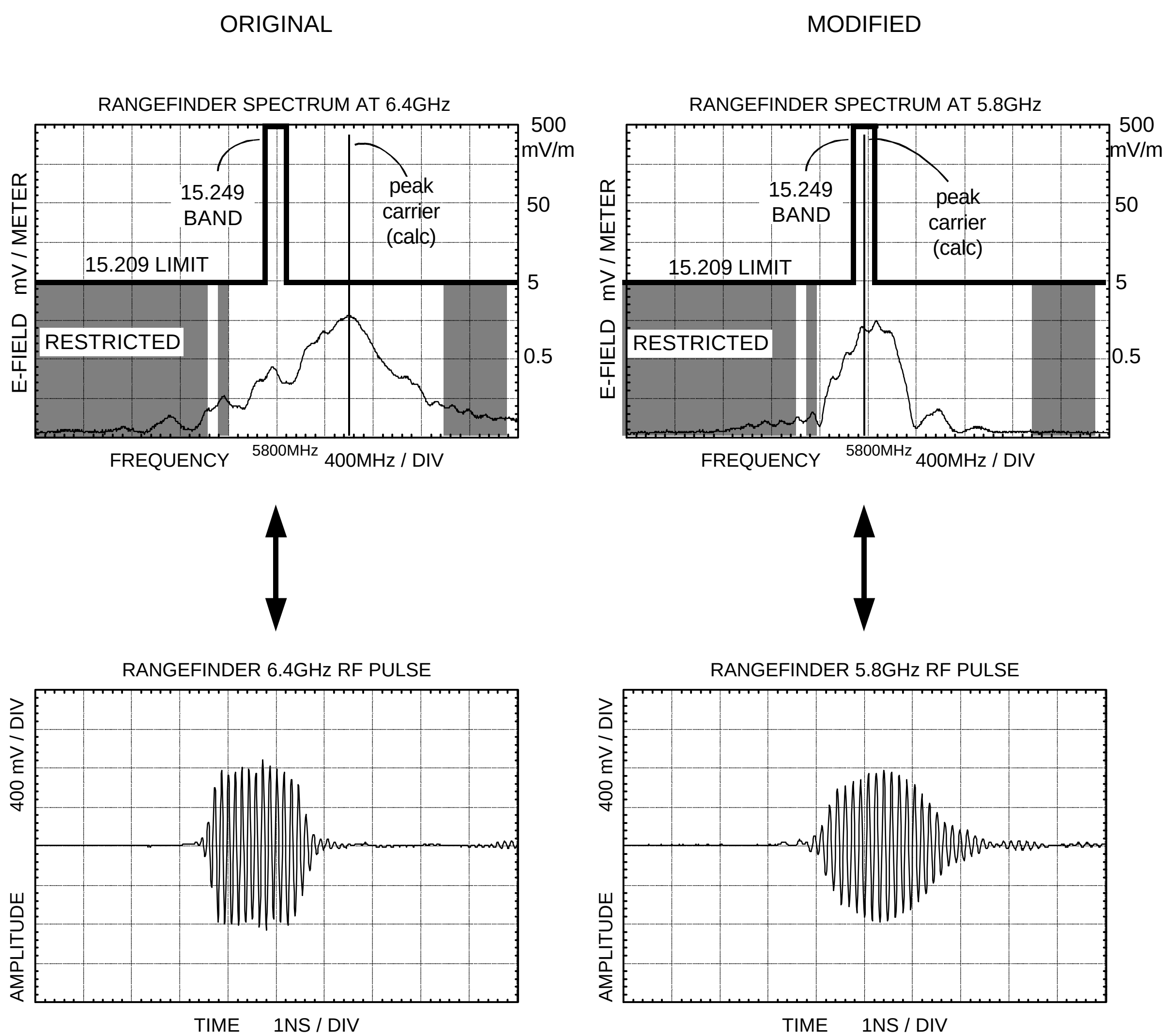
The oscillator is tuned to a carrier frequency within the 5725-5875MHz FCC Part 15.249 ISM band. Since the RF pulse generates a broad spectrum, the center frequency of the oscillator is difficult to deduce from a spectral plot. Consequently, the RF oscillator frequency was tuned by forcing it to oscillate continuously by applying DC bias to its drive node. The actual tuning of the oscillator to 5800MHz may be accomplished by adjusting the gate lead inductance, either by bending the gate wire towards the PCB ground plane, or by mounting a tuning screw close to the gate wire.

Depending on the GaAsFET and parasitics, 2nd and 3rd harmonics can vary from –20 to –40dBc. To help ensure FCC compliance for harmonic emissions, a coupled-line resonator pair was inserted between the drain and the antenna, and tuned to 5.8GHz. These resonators were also coupled to a 5.4GHz $\lambda/4$ trap to further reduce spurious emissions in the 5.4GHz restricted band. All the resonators were $\lambda/4$ elements comprised of 30 gage wire ~1.25cm long. They could be implemented in microstrip for production versions.

The plots below labeled ORIGINAL show the unmodified spectrum and corresponding time-domain waveform of the RF emission. The spectrum is centered at 6.4GHz and will only meet FCC 15.209 limits (after applying PDCF) by reducing the emission amplitude. The advantage to operation at 6.4GHz is (1) lower potential interference from other spectrum users (e.g., from ISM band users), and (2) relaxed center frequency tolerance, which could mean not having to tune the center frequency during production. While operation at 6.4GHz is legal in the U.S., it does require operating at a low emission level, and some long range radar applications may not tolerate “turning down the wick”.

By re-tuning to 5.8GHz, a much higher transmit power is permitted. The MODIFIED plots shown below depict the re-tuned emission spectrum centered at 5.8GHz and the shaped 2ns wide transmit pulse. The coupled-line resonators and 5.4GHz trap have clipped the wings of the spectrum and softened the RF pulse shape. Spurious emissions in the 5.4GHz restricted band have been reduced by more than 20dB below the emission peak. Both sets of spectral plots were taken using waveform averaging to effectively eliminate spectrum analyzer noise without altering the spectral shape.

The RF carrier stability is +/-20MHz over –55 to +65^{o}C. Since the RF carrier frequency must be 5800+/-75MHz, including initial tolerance and aging, this is adequate. The 2nd harmonic at 11.6GHz is –40dBc and the 3rd harmonic at 17.4GHz is below –30dBc (i.e., below the spectrum analyzer noise floor when the oscillator is connected directly to an HP8565 spectrum analyzer). A 3dB pad was added at the transmitter output to provide FCC compliance margin, since the peak carrier emission is close to the maximum permissible in the 15.249 band. In a dedicated design that would be integrated onto a single board, the transmit antenna gain could be reduced while increasing the receive antenna gain by the same factor, or the oscillator drive or its output coupling could be reduced.



FCC prescreening

The calibrated field strengths shown in the above plots were obtained at TEM Innovations using a calibrated spectrum analyzer and a $\lambda/2$ dipole antenna cut for 5.8GHz. The following steps were taken:

1. A peak spectrum analyzer reading of –78dBm was observed at 5800MHz with 1-meter separation between the rangefinder and the dipole. This dBm reading was converted to a voltage level V_{spec}

$$V_{spec} = .223(10^{(dBm/20)}) = 28 \text{ microvolts}$$

2. The electric field strength was calculated using the antenna K-factor for a dipole

$$K \sim 6/\lambda = 116/m \text{ for } \lambda = 5.2cm \text{ at } 5800MHz.$$

   Thus,

$$E_{1m} = K\,V_{spec} = (116/m)*28uV = 3.26mV/m \text{ at 1-meter range}$$

3. Field strength E_{1m} is scaled for 3-meters distance (the FCC specified range)

$$E_{3m}= E_{1m}/3 = 1.09mV/m$$

   Note that this level is well-below the FCC limit of 5mV/m for <10% duty cycle signals in the 15.209 bands. A 1-meter range was used instead of the FCC specified 3-meter range since the signal is fairly weak. This is common practice at the FCC test labs.

4. Calculate and apply PDCF. Given a spectrum analyzer bandwidth B = 1MHz and a pulse width T = 2ns,

$$PDCF = -20\log(1.5BT) = 50.5dB \text{ or } 333x \text{ in voltage.}$$

   The true peak carrier amplitude is

$$E_{3m}(carrier) = PDCF* E_{3m} = 333*1.09mV/m = 363mV/m$$

   The limit in the 5.8GHz band is 500mV/m for emitters with <10% duty cycle, so the rangefinder has some margin.

There may be some question about the accuracy of the PDCF, so data was taken using a sampling oscilloscope to measure the actual peak carrier voltage in the time domain. The results were within 1dB of the spectrum analyzer with PDCF.

Receiver

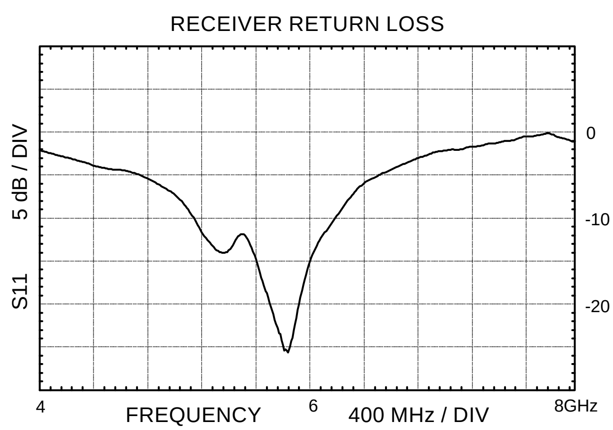
The LLNL demo unit receiver is a wideband sampling peak-hold circuit constructed in accordance with U.S. Patent 5,345,471 to McEwan. It was designed for use in a wide variety of applications including time domain reflectometry (TDR). As previously stated, it was not originally tuned for operation at 5.8GHz. With tuning, primarily in the form of input matching (see the RECEIVER RETURN LOSS plot), the –3db bandwidth was moved from ~5GHz to ~7GHz with about 2x more voltage response at 5.8GHz (see the RX BANDWIDTH plot). Several other minor changes resulted in a lower noise floor.

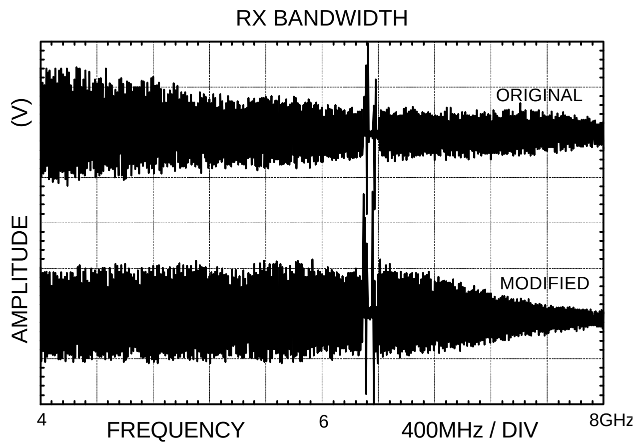
The RX RESPONSE TO 100uV 5.8GHz SINEWAVE plot (below) was taken with the receiver sampling at 1.25MHz with an RF input at the 4640th harmonic at 5800MHz, offset by 7kHz to produce the 7kHz beat frequency seen in the plot. Both traces represent overlays of about 100 sweeps with the vertical bars on each trace representing the peak-to-peak noise. An HP8672A microwave synthesizer provided the signals for this test.

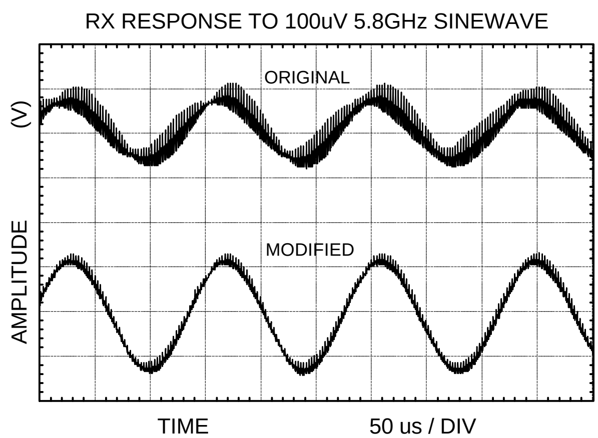
The noise floor of the modified receiver is about –95dBm. The receiver noise figure NF can be inferred as follows

$$-95\text{dBm} = -174\text{dBm} + 10\log(B) - 10\log(N) + NF$$

where –174dBm = KTB noise floor in 1Hz bandwidth, B = bandwidth ~ 7GHz, N = number of pulses averaged = $PRF/2\pi F_{BB}$ = 1.25MHz/($2\pi F_{BB}$) = 28.4, where PRF = pulse repetition frequency and F_{BB} = the detected baseband bandwidth = 466Hz (a 7KHz, Q = 15 bandpass filter was employed at the detector output). Thus NF = 6.7dB, a reasonable figure for a device without a low noise preamp. By adding a preamp with NF=1dB, the system S/N will improve by ~6dB.







Upgrade: The LLNL receiver uses a step recovery diode (SRD) to achieve greater than 5GHz bandwidth. TEM Innovations has a receiver that achieves 7GHz bandwidth using a low cost transistor instead of an SRD. Contact TEM for details.

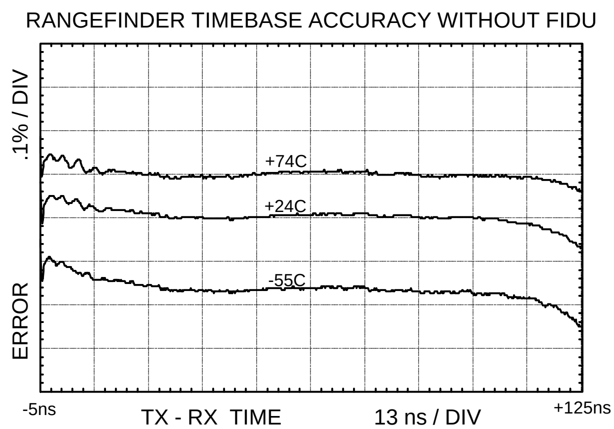
Timebase

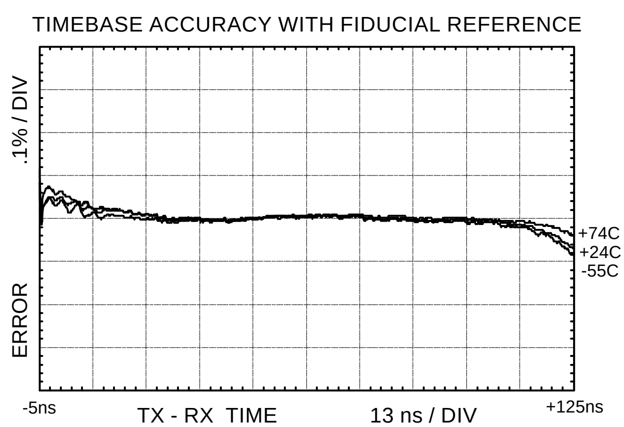
The precision timebase in the demo unit is a delay-locked-loop, or DLL, and is based on U.S. Patent 5,563,605 to McEwan. The reader is referred to this patent for a full description of the circuit.

The accuracy of the circuit, in terms of errors relative to an arbitrary full scale, was measured using a precision phase-to-voltage converter, or phase detector. While phase detectors are well-known, and many are capable of taking the measurements described below, TEM Innovations developed and optimized a version for it own use. A detailed circuit diagram of the TEM phase detector is available to LLNL MIR licensees and TEM licensees.

The plot below shows the accuracy achieved for TX-RX delays of –5 to + 125ns, expressed as a percent of full-scale error. Thus, 0.1% of 130ns full scale is 0.13ns or about 2cm (0.8") in range. The plot shows a very flat error curve but there is substantial drift with temperature in the vertical direction. This is timing offset drift, and not scale-factor or slope drift. Offset drift can be corrected with a circuit described in the patent. However, that circuit was eliminated in the demo unit since a main-bang fiducial referencing method is used to effectively eliminate offset drift.

The scale factor accuracy and stability of the DLL is set by the frequency of the quartz PRF oscillator, which is very stable, and by the ratio of two resistor pairs. A low-cost, 8-resistor precision tantalum nitride thin film resistor network was used to form the resistor ratios. Informal tests show the resistor ratios track with a stability of about 1ppm/^{o}C.





The plot immediately above is a graphical overlay of the first timebase plot (without fidu), showing the effect of a fiducial referencing system, as employed herein. There is essentially no change in slope with temperature (at the .01% level). Tests confirm that the main-bang fiducial reference system in the radar does cancel offset drift as expected.

In the fiducial referencing system, the range PWM is started by the main-bang pulse and stopped by the target return. Since both these pulses are time-locked on a single line, offset delays affect both pulses equally so there is no relative timing shift. Thus, large amounts of offset drift can occur in either the TX or the RX path with no effect on system accuracy. The primary sources of offset drift are logic gate propagation delay shifts in the TX and RX clock paths.

One possible enhancement to the LLNL DLL timebase is a digital coarse delay circuit that would provide delay increments of perhaps 20ns. The DLL would then provide high resolution interpolation within each 20ns increment. By using synchronous digital counters to generate the 20ns increments, the accuracy of the increments would be limited mainly by the frequency of the quartz PRF oscillator. Assuming a 0.1% accuracy DLL that interpolates over 20ns, the DLL interpolation accuracy would be .02ns or 3mm in range. Obviously, this hybrid approach could exceed .01% full scale accuracy.

> **Upgrade:** The LLNL timebase requires additional circuitry to eliminate the offset drift seen in the above plot. TEM has a novel DLL timebase that eliminates this drift source. TEM also has two synthesized timebases that rely on the accuracy of a quartz oscillator alone, with no resistor ratios involved. Contact TEM for details.

Signal processor

The signal processor (1) high- and low-pass filters the received signal, (2) scales the receiver gain in proportion to range (using a sensitivity time control or STC), (3) envelope detects the scaled received signal, and (4) threshold detects the TX and RX envelope-detected pulses using separate TX and RX automatic pulse detectors (called constant fraction discriminators or CFDs).

The default STC gain correction vs. range is set to increase receiver gain in proportion to radar range-squared. However, reflections from a large pool of water or oil require a correction proportional to range only, so one of the two variable gain stages was disabled by disconnecting the variable transconductance element in that stage.

The first order lowpass filter following the video absolute value circuit was increased to a second order filter by adding another RC stage. This eliminated a slightly jagged detection envelope without slowing the true RF envelope appreciably. In a dedicated version, an active third order Bessel LPF would be desirable.

Several time constants in the pulse detector circuit needed to be increased to accommodate the slower scan rate (9.5Hz) that was selected for 62-foot operation. This includes (1) the CFD peak hold capacitors, (2) the reset pulse was stretched to inhibit any potential glitch detection for the first few percent of the range sweep, and (3) a minimum TX-RX interval time constant was set with an RC network (this prevents the receive detector from trigging on the TX pulse). The TX CFD flip-flop was intermittently double triggering, so its input filter rolloff was increased to eliminate this problem.

The NEAR GAIN and FAR GAIN potentiometers must be adjusted to ensure that the main bang pulse is always the largest pulse (or within a factor of two of the largest) to appear at the system output. Otherwise, an overly large target return pulse may capture both the TX and RX CFDs. Once the gain controls were properly set, reliable detection was observed.

For applications with rapidly moving targets (e.g., wavy water), the CFDs should be replaced with a memory-free type automatic detector, since the peak detectors in the CFD are unable to follow rapid fluctuations in pulse amplitude.

The performance of the detection circuit was evaluated using a simulated pair of pulses derived from a Tektronix PG508 pulse generator and tabulated below. The simulated pulses had the same risetime, falltime and width as the actual pulses. Since the return pulses should always be within 0.3-3V, pulse detection errors are seen to be less than 2mm.

Pulse Detector Tracking Error

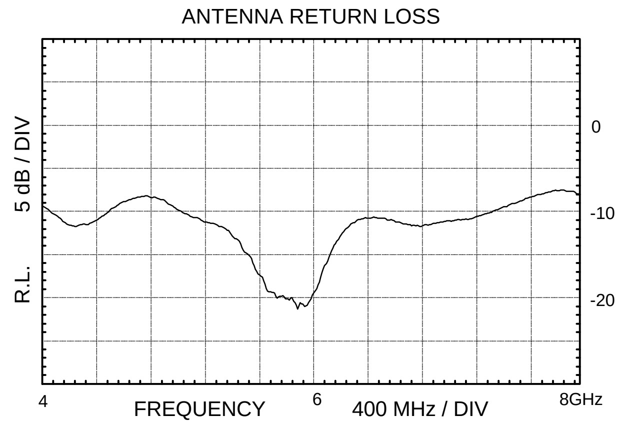
Pulse Amplitude	**ΔT**	**Δ-range error**
3V	0	0
1.0	0	0
0.3	0	0
0.1 (under range)	10us	2mm (.075”)
.03 (under range)	30us	6mm (.225”)

Upgrade: The LLNL system requires main-bang coupling for offset correction. TEM has developed a high accuracy technique that eliminates main-bang coupling so close-in returns are not obscured and the minimum target range can approach zero. Contact TEM for details.

Antennas

The modular system antennas are based on U.S. Patent 5,754,144 to McEwan. They are horns with a specialized UWB feed structure and are not particularly optimized for 5.8GHz operation. The damping resistors used for UWB operation were deleted and a –20dB return loss at 5.8GHz was obtained (99% radiation efficiency, neglecting slight losses). Small dielectric slabs (.062” thick FR-4) were added to the feed structure to account for some manufacturing misalignment and to better match the antennas at 5.8GHz. This design is not recommended for 5.8GHz work; a generic 5.8GHz horn with an E-field probe will work as well with reduced alignment requirements.

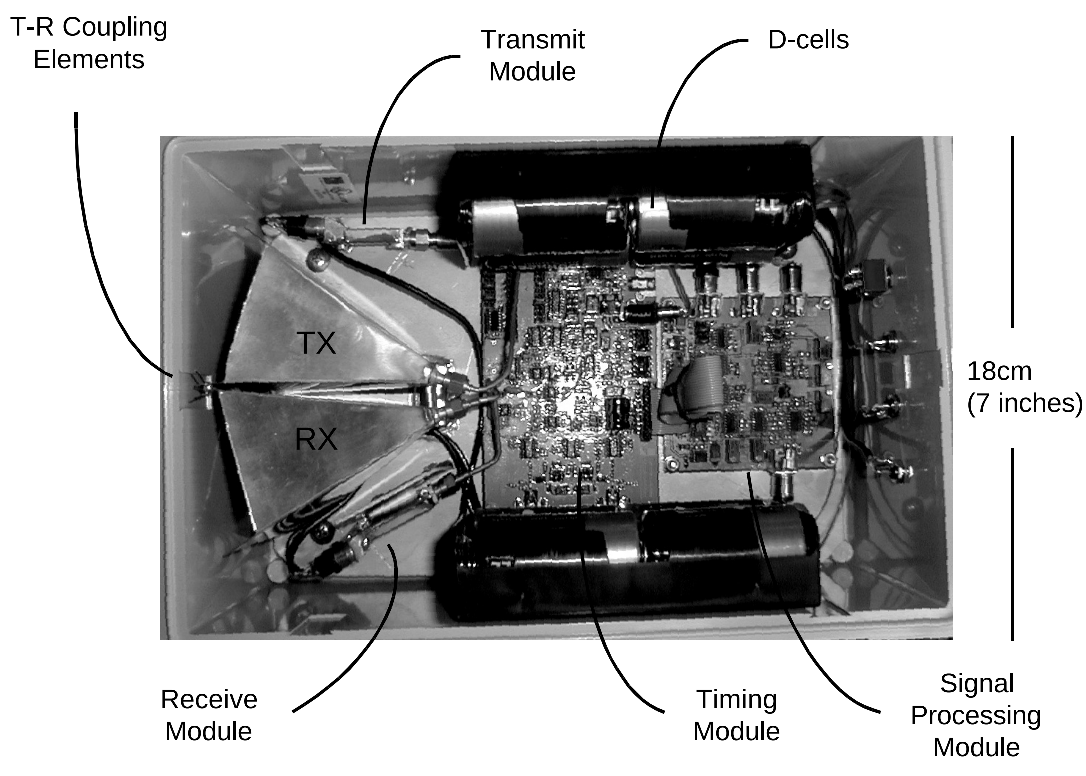
The calculated gain is 13dBi, and the measured system azimuth beamwidth is 32 degrees full width at the –3dB points.



> **Upgrade:** TEM has a complete rangefinder on a 6.25x10cm (2.5x4”) PCB including simple PCB-mounted horn antennas and operation at either 5.8 or 24GHz. Contact TEM for details.

Packaging

The demo unit was housed in a weather-tight plastic NEMA box as shown below. For demonstration purposes, alkaline D-cells were used to provide about 100 hours of service at a current drain of 150mA. The range readings settle to 0.1 percent of F.S. within a few seconds after power-up, so low duty cycle measurements are possible.



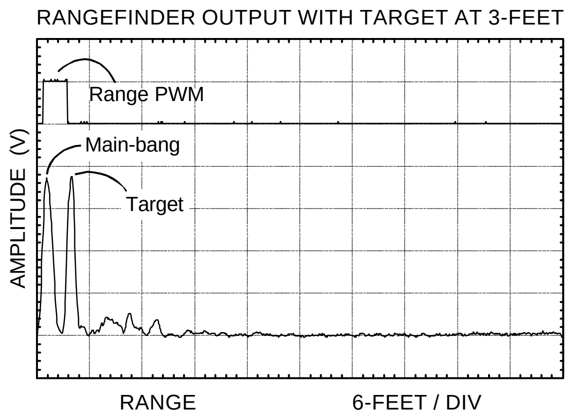
The horn antennas were mounted off-axis since this particular design has a fair amount of beam squint at 5.8GHz. The arrangement shown in the photo provides optimum gain. The main-bang coupling was not difficult to control—an acceptable amount of coupling was provided with almost any arrangement of the horns. Two small $\lambda/4$ wires labeled "T-R coupling elements" in the photo were added to buck a small tendency for the RF pulse to ring and stretch the main bang pulse slightly. In a production design, the effect of the plastic wall in front of the horns can be effectively eliminated by proper spacing of the horn-to-wall distance.

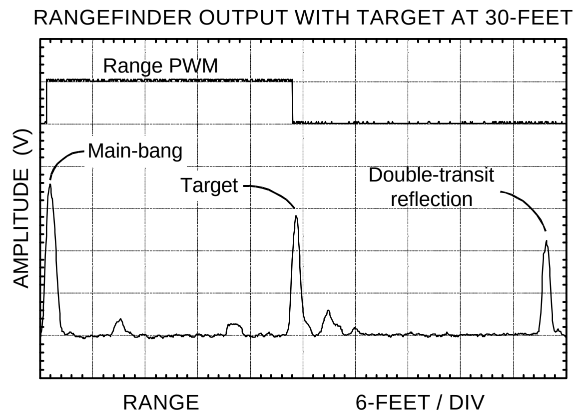
System Tests

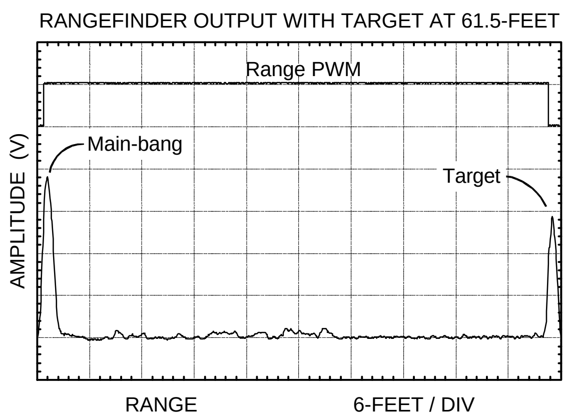
1. Range tests

The rangefinder was placed on a bench and pointed at a $0.7m^2$ flat metal target plate, which simulated a liquid surface. The upper left plot shows the main-bang, target return, and range PWM at a 3-foot range. The upper right plot shows the same outputs for a 30-foot range. There is some clutter on the baseline between the main-bang and the target return due to a substantial amount of physical clutter (cardboard and metal boxes, etc) only a few feet off the beam line. A double-transit reflection appears at a 60-foot range due to the metal target reflecting back to the metal bench that the rangefinder was mounted on, and then back to the target and back to the radar. This reflection is not a problem since it occurs after the target is detected, which is the first pulse to be detected after the main-bang.

The lower plot shows a return at 61.5-feet. Note that the clutter baseline has changed since the radar had to be relocated to shoot through a garage door to a target outside the test facility. The amplitude of the target pulse is the essentially the same at all ranges due to the action of the STC. In practice, the clutter environment must be taken into account for high accuracy rangefinding.







2. Temperature dependence

In situ temperature dependence tests were run at three different ranges. Since an environmental chamber was unavailable, a heat gun was pointed inside the NEMA box and the temperature was monitored with a thermocouple at one of the circuit boards. For cold temperature tests, difluororethane ("circuit freeze") was vigorously sprayed over the all the components in the NEMA box and readings were taken immediately after a white frost formed (-55^{o}C). The PCBs were conformally coated with a silicone compound to eliminate the effects of moisture condensation during the temperature tests.

Rangefinder Temperature Dependence

Temperature	**3-foot range (0.9m)**	**30.2-foot range (9.2m)**	**61.5-foot range (18.7m)**
-55^{o}C (-67^{o}F)	-.02% of FS	+.09% of FS	-.06% of FS
24^{o}C (75^{o}F)	0 (reference temp)	0 (reference temp)	0 (reference temp)
74^{o}C (165^{o}F)	+.09% of FS	+.06% of FS	+.09% of FS

The primary source of drift appears to be a shift in the shape of the RF pulse rise envelope and inadequate pulse detector tracking thereof. Temperature dependent oscillator turn-on characteristics are common in GaAsFET oscillators and can be minimized by design. Alternatively, the pulse detector tracking characteristics could be further explored.

Timing offset drift, scale factor drift, and pulse detector drift all appear to contribute less than 1/3 of the total observed drift—the bulk of the drift was localized to the transmit oscillator. Nevertheless, the maximum observed drift of .09% is within expectations for the LLNL timebase alone. Considering the complete system described herein, the timebase is performing beyond the 0.1% stability level.

3. Output scale factor

The pulse detector output, or PWM output has a range equation of

Range (PWM reading + 0.19ms)*7.5inches/ms

The 0.19ms offset could be removed with an RC time constant in the PWM output circuit. It could also be accounted for after data acquisition, which would be the preferred method since the scale factor must be calibrated in any case. Both the offset and scale factor must be calibrated during manufacture before the unit can be considered to be absolutely accurate.

4. Jitter and resolution.

Since the system has a 9.5Hz reset rate, range (PWM) readings are provided every ~0.1 second without averaging. A simple counter was used to measure the PWM width and provide averaging. With 10 averages, data is updated every second, and every 10 seconds with 100 averages. With 10 averages the peak jitter was less than 1 count at 3ft range, 2 counts at 30ft, and 3 counts at 61.5ft range.

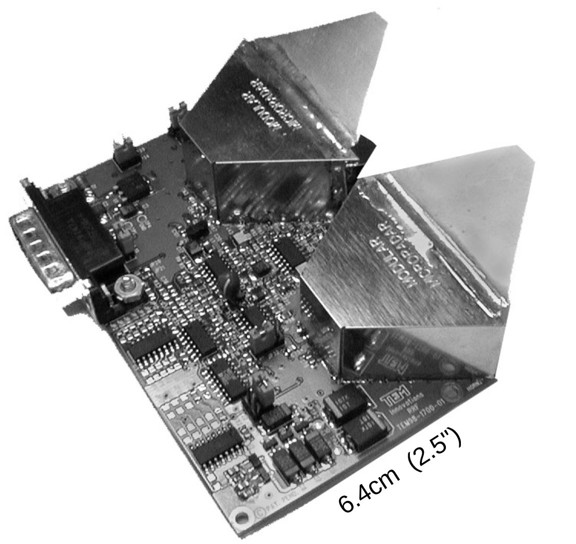
The jitter can be taken to be the minimum discernable resolution of the system, and 2 counts corresponds to 1/8” or 3mm. By averaging 100 readings the jitter would be reduced 10x, to 2/3 counts at 30ft, or .04” (1mm). Thus, the resolution is about .04” with averaging.

Summary

While the generic LLNL demo unit was not designed for a specific application, the results presented here support the overall flexibility and accuracy of the technology. We have shown that the demo unit rangefinder meets the following

- 2 to 62-foot operational range (0.6 to 18.9-meters)
- better than 0.1% of full scale timing accuracy over temperature
- 1mm range resolution
- FCC Part 15 compliance prescreening at 5.8GHz.

Better results with fewer components and a more compact design can be realized with a dedicated design on a single printed circuit board, as presently implemented with TEM Innovations upgrades that rely on TEM intellectual property.



The TEM Innovations 5.8GHz precision rangefinder Model RRF58.