IMPLEMENTATION OF RF CIRCUITRY FOR REAL-TIME DIGITAL BEAM-FORMING SAR CALIBRATION SCHEMES

Stephen J. Horst, James P. Hoffman, Dragana Perkovic-Martín, Scott Shaffer, Tushar Thrivikraman, Phil Yates, and Louise Veilleux
Jet Propulsion Laboratory, California Institute of Technology
4800 Oak Grove Dr., Pasadena, CA 91109, USA
sjhorst@jpl.nasa.gov

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Abstract

The SweepSAR architecture for space-borne remote sensing applications is an enabling technology for reducing the temporal baseline of repeat-pass interferometers while maintaining near-global coverage. As part of this architecture, real-time digital beam-forming would be performed on the radar return signals across multiple channels. Preserving the accuracy of the combined return data requires real-time calibration of the transmit and receive RF paths on each channel. This paper covers several of the design considerations necessary to produce a practical implementation of this concept.

1 Introduction

The SweepSAR architecture promises significant performance improvements over traditional strip-map SAR instruments for remote sensing applications [1]. This technique produces an extremely wide swath on transmit, while selectively sweeping the return signal on receive using multiple real-time digital beam-forming (DBF) channels. This wide-swath capability is critical for reducing the temporal baseline of repeat-pass interferometer instruments, which plays a critical role in enabling Interferometric SAR (InSAR) deformation measurements for emergency response applications after an earthquake. SweepSAR also increases the number of looks compared to the ScanSAR architecture, an alternative wide-swath technique in which a single narrow swath is scanned in time over a broader area. Comparative analysis has shown that the SweepSAR architecture is estimated to reduce mass by 70% and costs by 50% over a traditional phased-array design, such as ScanSAR, by reducing the number of necessary T/R channels [4].

Implementation of a real-time DBF system requires the gain and phase of individual RF channels to be well understood for both transmit and receive activities [3]. Unaccounted shifts and uncertainty in the RF front-end will corrupt the digital algorithms that synthesize the return signals. Figure 1 shows the basic architecture of the RF portion of the radar. On transmit, mismatches between channels lead to increased range ambiguities caused by increased variation in the transmitted wave-front, which impose requirements that gain and phase be controlled between channels to within a specified tolerance. On receive, the time-critical gating of the receiver channels will produce distortion of the return signal without real-time knowledge of the gain and phase of the channel. This distortion leads to degradation of the integrated and peak side-lobe ratios. Each of these factors will play a role in the uncertainty of a deformation measurement.

Traditional SAR calibration techniques rely on post-processing to remove artifacts of the RF front-end in producing the image. This process involves pre-flight characterization of the RF module performance over temperature, followed by the application of correction factors in post-processing based on in-flight temperature measurements synchronized to the science data. Effects not related to temperature, such as radiation effects and aging, are not accounted for using traditional calibration. Furthermore, the need for real-time DBF requires that correction factors be applied on-board rather than in post-processing. On-board correction is a fundamental requirement because the DBF algorithm is essentially a weighted sum of the complex returns within a group of receive channels. Once the summing operations have been performed, it becomes impossible to untangle the result to apply correction factors to
Designing RF hardware that can reliably measure Solid Earth deformation on the scale desired by the science community requires understanding how gain and phase errors in the hardware will impact the desired result. A complete derivation of radar performance for this purpose is given in [2], with the relevant equations reproduced here for convenience. Uncontrolled variance in the phase of the instrument will degrade the integrated side-lobe ratio (ISLR) metric. If we start by assuming the random fluctuations in the phase estimates have a normal distribution, which may be a reasonable assumption, the ISLR can be expressed as the RMS value of the various phase noise sources

$$ISLR = 20 \log \left( \sqrt{\sigma^2_{RX} + \sigma^2_{TX} + \sigma^2_{ADC}} \right)$$  \hspace{1cm} (1)$$

These will combine with the total ambiguity-to-signal ratio ($AMBR$) and quantization-to-noise ratio ($QNR$) to give a total estimate of the multiplicative noise ratio ($MNR$)

$$MNR^{-1} = ISLR + AMBR + \frac{1}{QNR}$$  \hspace{1cm} (2)$$

The multiplicative noise can be combined with the desired signal to noise ratio ($SNR$) in linear terms to give the correlation of measurements of the scatterer

$$\gamma = \frac{SNR}{SNR(MNR^{-1} + 1) + 1}$$  \hspace{1cm} (3)$$

Finally, the deformation variance can be determined by combining the correlation of the measurements with the number of looks used to produce the final estimate to give

$$\sigma^2_{deformation} = \frac{\lambda}{4\pi} \sqrt{\frac{1 - \gamma^2}{\gamma^2}}$$  \hspace{1cm} (4)$$

These equations give a translation between a phase error in the instrument, $\sigma^2_{RX}$ and $\sigma^2_{TX}$, to an error in the measured deformation at the target, $\sigma_{deformation}$. For the purposes of this analysis, we will target a deformation error less than 0.2 mm and neglect $AMBR$ and $QNR$, while allowing $SNR$ to go to infinity. If we split the variance of the TX chain and RX chain evenly, while disregarding any variance of the ADC, we have a means of observing the impact of hardware phase variance on deformation measurement accuracy in the absence of other factors. A plot under these assumptions is shown in Figure 2. For resolving changes down to 0.2 mm, this implies a total phase variance of less than $\pm0.05^\circ$ under ideal circumstances.

Phase variation in the T/R module hardware can occur from both random and deterministic sources. Random noise sources fit easily into the ISLR estimation in (1). However, systematic errors primarily due to temperature variation can degrade the measurement if left unaccounted. The on-board calibration process eliminates a majority of these systematic variations in phase. However, the components outside the calibration loop or those that make up the calibration circuitry themselves remain susceptible to these errors. For these components we can derive certain requirements to aid in the design. For the components making up the calibration paths, this implies keeping the total phase deviation over temperature less than the $\pm0.05^\circ$ requirement through the use of a look-up table. By restricting those out-of-loop components to relatively stable items such as couplers and switches, a single temperature sensor should be accurate enough to achieve this distribution, as demonstrated later in Figure 5.

A more interesting constraint arises from the isolation of signals through the various paths that make up the on-board calibration. The bleedthru from an isolated signal path can be modeled as an interferer on top of an ideal signal

$$e^{i\omega t}$$

where $M$ and $\theta$ are constant terms for the amplitude and phase of the interferer respectively. The reference term $e^{i\omega t}$ is given here as a sinusoid but can also be replaced with a linear chirp where appropriate. The relative strength and phase of the interferer will determine the amount of absolute phase error observed in the combined signal through the vector addition of the first term in (5). However, absolute phase changes do not impact the repeat-pass interferometer, only a difference in phase. Therefore, it is the slope of the vector sum with respect to phase that is important here. This can be written as

$$\frac{1}{1 + Me^{i\theta}}$$
Figure 3: Measured phase delta relative to the phase offset of an interferer. In order to keep systematic phase variance within an acceptable range, the interferer magnitude (isolation) must be suppressed by at least 55 dBc.

\[
\frac{d}{d\theta} \left(1 + M e^{j\theta}\right) = \frac{d}{d\theta} \tan^{-1}\left( \frac{M \sin \theta}{1 + M \cos \theta} \right)
\]

Figure 3 shows the result of (6) for interferers of various magnitudes, \(M\), swept across all possible phase offsets, \(\theta\). The results show that in order to keep change in measured phase below an acceptable level for any possible phase offset, the magnitude of the interferer (set by the isolation between paths in the T/R module) must be suppressed by at least 55 dBc from the desired signal. The alternative is to control the phase relationship of any possible interferers to within a narrow window, a risky and difficult task. From Figure 3, if we allow the interferer magnitude to increase to -40 dBc, we must control the phase to within \(\pm 10^\circ\) of either 90\(^\circ\) or 270\(^\circ\). Understanding these constraints allow for informed design decisions in the T/R module.

3 T/R Module Design

The reconfigurable architecture for a single T/R channel employing the proposed closed-loop digital calibration scheme is shown in Figure 4. There are three primary switch-configurable paths for characterizing the gain and phase performance of the T/R module: the transmit calibration path, the receive calibration path, and a bypass calibration path that has separate sub-configurations for transmit and receive because of the anti-alias filter.

We discuss receive calibration and transmit calibration in separate sections below. Each of these uses a bypass calibration path as a reference to isolate the impact of the desired calibration path. Functionally, this path is used for correcting skew between T/R channels, which is manifested either by phase offsets in the common RF paths, or clock skew between the digitizing ADCs in each channel. However, performing a bypass calibration comes at a cost to the science data. The wide swath covered by the SweepSAR transmit pulse stretches the return signal over a longer period of time. In fact, the receive window is so long that valid return data is constantly incident somewhere on the receive array. This implies that the act of transmitting itself causes a loss of data. The bypass calibration is a similar scenario where the receiver must be turned off in order to perform the calibration. Therefore, minimizing the number of times a bypass calibration needs to be performed is critical to maximizing the performance of the instrument.

The placement of the anti-alias filter has significant impact on the bypass calibration scheme. In our design of an L-band SweepSAR instrument, the radar return data is directly sub-sampled at a rate of 240 Msps. The desired 80 MHz bandwidth is thus folded over several times in the digitization process. The anti-alias filter is necessary to suppress the out-of-band noise from folding on top of the desired signal and reducing the SNR. At the same time, a filter is needed on the transmit calibration path to prevent the harmonics generated from the highly compressed HPA from corrupting measurements of the transmit chirp. Rather than using two large filters, we have chosen to use a single filter just before the ADC. The impact on the bypass calibration is that we must now use two separate paths for TX and RX bypass as shown in Figure 4. By including the filter in the TX bypass measurement, we can remove its impact when measuring the TX chirp for calibration. However, if we were to use this same path for the RX bypass, we would also remove the impact of the filter during RX calibration even though the filter is very much part of the desired RX path. Since the filters are one of the most volatile components in the T/R module over temperature, this is not acceptable. The additional bypass path that excludes the filter is therefore used for RX calibration so that the filter’s impact will be included in the RX correction algorithm.

The introduction of a split bypass path also creates another loop that must meet the isolation requirements developed in section 2. This path travels across the throws of SW2, across the outputs of SP1, and through the isolated path of SW3. If we assume a Wilkinson power divider achieves 15 dB of isolation between the output ports across the band, then this implies that the switches SW2 and SW3 need at least 20 dB of isolation each in order to meet our -55 dBc isolation requirement.
The gain and phase information extracted from the cal tone is used to adjust the weighting factors applied by the DBF algorithm.

Phase stability measurements of a development receiver built to test the calibration concept were taken over temperature, and are shown in Figure 5. After stabilizing at each temperature point, 100 measurements were taken over the course of fifteen minutes on the network analyzer to get a sense of the circuit’s variation over an extended period of time. The mean value of these measurements is plotted in the figures with error bars representing a ±3σ variation based on 100 samples at center band. The variation seen across the 100 measurement window is very strongly correlated to temperature although the overall phase response deviates from linearity at lower temperatures. The results of this test give an indication of the random variance that can be expected from the receiver as a whole. In section 2, we established that the cumulative phase variance, σ^2_{RX+TX}, had to be less than ±0.05° in order to meet the desired deformation measurement accuracy. Measurements from Figure 5 indicate that the receiver contribution to that, σ^2_{RX}, at a fixed temperature is orders of magnitude less than that. This is important because it establishes that systematic variances that can be corrected in calibration will dominate the allocated error margin as opposed to random fluctuations which cannot.

While performing receiver calibration, the AWG in Figure 1 that typically generates the transmitted chirp will instead generate the CW calibration tone. This will follow through the RX cal path until it is combined with the radar return signal at CP2 in Figure 4. Isolating the bypass path to the required level means all three switches have isolation better than 20 dB, identical to the requirement for the bypass path developed earlier. There are two paths through the TX chain that need to be sufficiently isolated using this scheme: the TX calibration path and the reflection off of the antenna and back through the un-calibrated circulator. The large attenuation needed in the TX calibration path to bring the +50 dBm transmitted waveform to a level easily handled by the ADC means that isolation through this path is easily sufficient to meet the requirement. The path reflected off of the antenna is not as easy.

It is desirable for the cal tone injected into the receiver at CP2 to be at a level comparable to the expected radar return signal. This means the signal should be no greater than around -60 dBm, and the requirement to be -55 dBc isolated will be relative to that. If the AWG signal enters the T/R module at a level of 0 dBm, then the isolation between the T/R module input and the input port of CP2 needs to be greater than 115 dB. Part of this comes from powering down the TX chain during receive. However, because of the switching speed required between TX and RX events to reduce the loss of data, not all of the amplifiers can be turned off. The high power parts naturally lend themselves to high-speed switched operation but the small signal amplifiers that have the most gain cannot turn on and off fast enough for this application. That is the reason SW1 is an SP3T switch in-line with the TX chain. A coupled switch configuration with an in-line coupler followed by the switch out of the transmit path would provide a more stable path for the transmit calibration but would allow the RX calibration tone to be amplified directly by the small-signal amplifiers of the TX chain, which would fail the isolation requirement by over 20 dB.

### 3.1 Receiver Calibration

Receive calibration is implemented using one or more CW tones generated just outside of the chirp bandwidth, but within the receive filter bandwidth. This signal is coupled in with the radar return signal whenever the receive gate is open [3]. Digital filtering is used after the signal is digitized to separate the calibration tone from the return data. The gain and phase information extracted from the cal tone is used to adjust the weighting factors applied by the DBF algorithm.

Figure 5: Phase response of the receiver over temperature at center band. Phase measurements were taken 100 times over a 15 minute period.

### 3.2 Transmitter Calibration

Transmit calibration consists of a coupled version of the transmit chirp directly digitized by the ADC. This signal is used to maintain phase alignment of the transmit channels by making real-time adjustments to an internal phase shifter in each channel and is also sent to the ground for correcting residuals in post-processing. Transmit calibration occurs every time a chirp waveform is transmitted. The bypass calibration includes the anti-alias filter as mentioned previously so that it can be removed from the calibration, since it is not part of the true TX chain. Keep in mind the absolute phase transmitted by each channel is not important. It is only important to keep the phases of each channel aligned relative to each other to within a few degrees. The bypass calibration is therefore included so that phase shifts in common blocks such as the AWG and up-converter do not cause all of the T/R module phase shifters to change at once, improving the reliability of the transmit calibration.

The decision to combine the filter needed at the output of the HPA with the anti-alias filter reduces the size footprint of two large components into one large component. More importantly, it directly improves the SNR of the instrument by not including a filter directly in the transmit path. With each transmitter outputting 100 W the loss of even a few
cable lengths of at least 80 inches will be needed, which be distributed to each of the twelve proposed T/R modules, array element spacing. For the single L-band chirp signal to signals arise from the physical distances mandated by the T/R module at a later time. These re-reflected, or “ghost”, through the transmission line, then reflect back again off interfaces, a portion of the desired signal will reflect back module on a single channel. Due to mismatches at the in-

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3.3 True Time Delay Interference

So far the impact of isolation has been analyzed for the re-

configurable paths within the T/R module. One final iso-

lation consideration arises from the distribution network to

the T/R modules. Referring to Figure 1, this occurs because of the time delay introduced by the long phase matched ca-

bles extending from the distribution network to the individ-

ual T/R modules. This time delay can also be modeled as an interfering signal as developed in section 2 as long as the signal is a CW sinusoid. However, since the cause of the phase shift is a true time delay, a linear chirp needs separate consideration.

The problem can be highlighted in Figure 6a, showing the interface between the distribution network and the T/R module on a single channel. Due to mismatches at the in-

terfaces, a portion of the desired signal will reflect back through the transmission line, then reflect back again off of the distribution network interface to arrive back at the T/R module at a later time. These re-reflected, or “ghost”, signals arise from the physical distances mandated by the array element spacing. For the single L-band chirp signal to be distributed to each of the twelve proposed T/R modules, cable lengths of at least 80 inches will be needed, which introduce a delay of 8.4 ns through each cable. Poor return loss from the T/R module and distribution network coupled with the delay introduced by the cable length could easily produce coherent ghost signals that exceed the -55 dBc isolation requirement.

Our solution to this problem has been to add attenuators to the interface between the T/R module and the transmit distribution network. As shown in Figure 6b, these attenuators reduce the magnitude of the ghost signal at the expense of requiring a stronger driver amplifier at the input of the transmit distribution network. Two approaches were considered. The first places all of the attenuation at the T/R module interface, while the second splits the total attenuation in two with each half split between the T/R module and distribution interfaces respectively. The impact of the attenuators on the magnitude of the ghost signal at the T/R module input is shown in Figure 7. In order to meet the 55 dB isolation requirement for the ghost signal, a distributed attenuator approach with 16 dB total attenuation is desired. However, this would require an amplifier with +31 dBm output power to drive the transmit distribution network. The amplifier in the present distribution network falls just short of that power level at +28 dBm. Meeting the isolation require-

ment with the present driver amplifier requires improving the VSWR of the interface components to better than 1.1:1. For a sinusoidal signal, the impact of the ghost re-reflection will appear identical to the analysis in section 2 because the frequency is constant over time. However, the frequency of a linear chirp is changing over time. This actually reduces the impact of the interferer because the reference term from (5) requires proper alignment of the chirp to maximally detect the signal. With the ghost signal delayed in time, it is
no longer making a maximum disturbance on the desired signal. The result of a delayed chirp simulation is shown in Figure 8. The total delay is estimated for an 80 inch cable with a phase velocity of 80%. The swept phase offset is the result of considering all possible phase angles off of reflections at the interface components. Under these conditions, the isolation requirement for the time delayed chirp drops to -40 dBc. From Figure 7a, we can see that this requirement can be met without the use of the attenuators needed in the CW case.

4 Summary

The need to perform real-time digital beam-forming on a SAR instrument produces unique design challenges to accommodate the necessary on-board calibration. We have outlined our efforts to design a re-configurable T/R module that would meet the level of precision desired by the Solid Earth science community. These steps have included setting bounds for the permissible phase variation in the T/R module and establishing the isolation required through each of the re-configurable paths. We have also established that systematic variations such as those caused by temperature fluctuation will dominate over random phase variations in the actual hardware. Finally, efforts to reduce interference caused by the time delay from the distribution network to each T/R module have been presented. This effect is dulled by chirped waveforms as the delay offsets the frequency with the desired signal to reduce the impact on the phase estimate at the ADC.

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References


