5. RVP7 Processing Algorithms

This chapter describes the real-time data processing algorithms implemented within the RVP7 signal processor. The discussion is confined to the mathematical description of these algorithms. Figure 5–1 shows the overall process by which the RVP7 converts the IF signal into corrected reflectivity, velocity, and width. Table 5–1 summarizes the quantities that are measured and computed by the RVP7. The type of the quantity (i.e., real or complex) is also given. Subscripts are sometimes used to denote successive samples in time from a given range bin. For example, s_n denotes the "I" and "Q" video sample from the n'th pulse from a given range bin. In cases where it is obvious, the subscripts denoting the pulse (time) are dropped. The descriptions of all the data processing algorithms are phrased in terms of the operations performed on data from a single range bin — identical processing then being applied at all of the selected ranges. Thus, there is no need to include a range subscript in this data notation.

It is frequently convenient to combine two simultaneous samples of "I" and "Q" into a single complex number (called a phaser) of the form:

$$s = I + jQ$$

where "j" is the square root of –1. Most of the algorithms presented in this chapter are defined in terms of the operations performed on the "s"'s, rather than the "i"'s and "q"'s. The use of the complex terms leads to a more concise mathematical expression of the signal processing techniques being used. In actual operation, the complex arithmetic is simply broken down into its real-valued component parts in order to be computed by the RVP7 hardware. For example, the complex product:

$$s = W \times Y$$

is computed as

$$Real\{s\} = Real\{W\}\ Real\{Y\} - Imag\{W\}\ Imag\{Y\}$$

$$Imag\{s\} = Real\{W\}\ Imag\{Y\} + Imag\{W\}\ Real\{Y\}$$

where "Real{}" and "Imag{}" represent the real and imaginary parts of their complex-valued argument. Note that all of the expanded computations are themselves real-valued.

In addition to the usual operations of addition, subtraction, division, and multiplication of complex numbers, we employ three additional unary operators: "||", "Arg" and "*". Given a number "s" in the complex plane, the magnitude (or modulus) of s is equal to the length of the vector joining the origin with "s", i.e.

$$|\, s \,| = \left(Real\{s\}^2 + Imag\{s\}^2\right)^{1/2}$$

The signed (CCW positive) angle made between the positive real axis and the above vector is:

$$\measuredangle = Arg\{s\} = \arctan\left[\frac{Imag\{s\}}{Real\{s\}}\right]$$

where this angle lies between $-\pi$ and $+\pi$ and the signs of Real{s} and Imag {s} determine the proper quadrant. Note that this angle is real, and is uniquely defined as long as |s| is non-zero. When |s| is equal to zero, Arg{s} is undefined. Finally, the "complex conjugate" of "s" is that value obtained by negating the imaginary part of the number, i.e.,

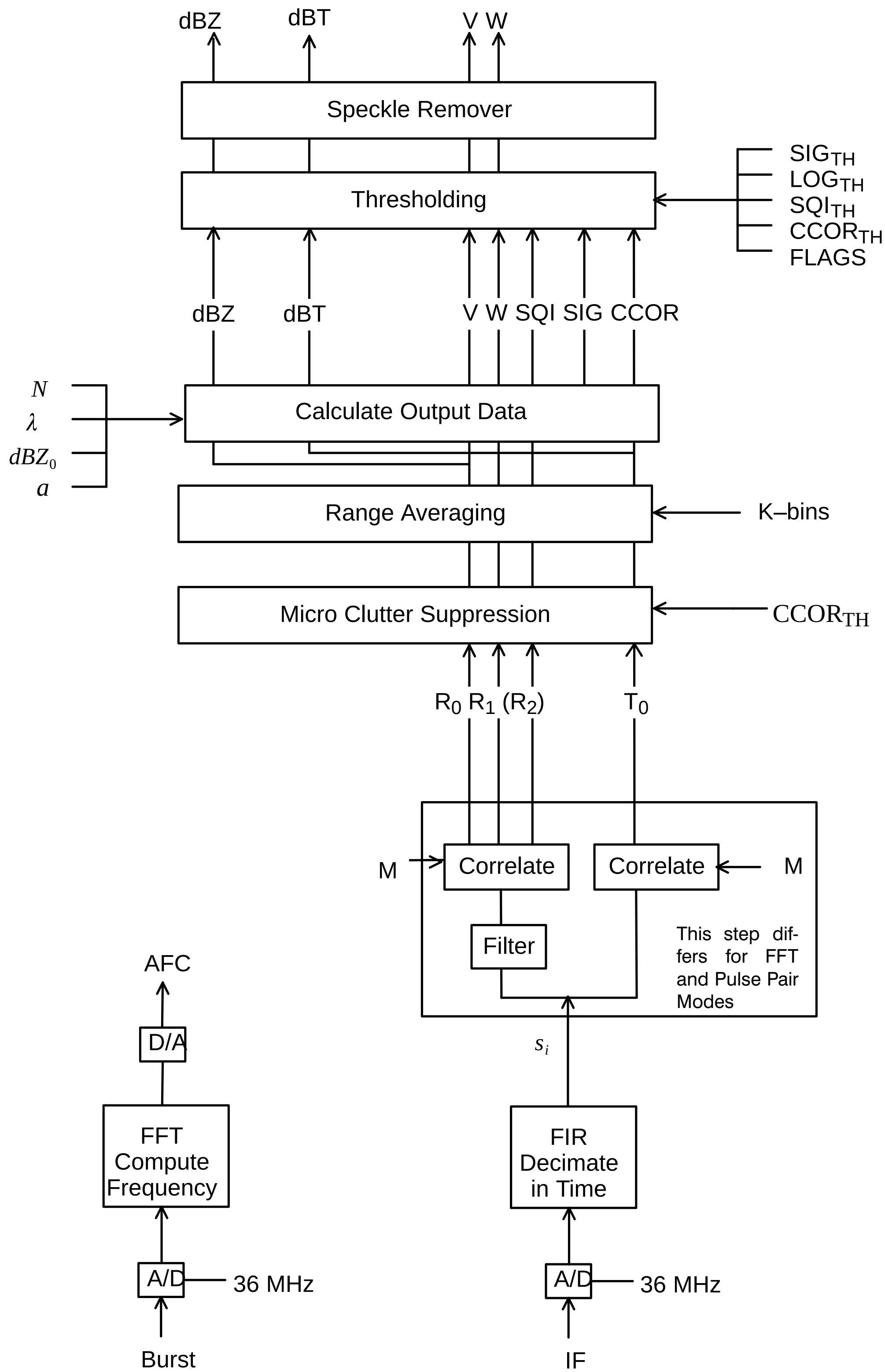
$$s^* = Real\{s\} - j\ Imag\{s\}.$$

Note that Arg{s*} = –Arg{s}. The reader is referred to any introductory text on complex numbers for clarification of these points.

Table 5–1: Algebraic Quantities Within the RVP7 Processor

p	Instantaneous IF-receiver data sample	*Real*
b	Instantaneous Burst-pulse data sample	*Real*
I, Q	Instantaneous quadrature receiver components	*Real*
s	Instantaneous time series phaser value	*Complex*
s'	Time series after clutter filter	*Complex*
T_0	Zeroth lag autocorrelation of A values	*Real*
R_0	Zeroth lag autocorrelation of A' values	*Real*
R_1	First lag autocorrelation of A' values	*Complex*
R_2	Second lag autocorrelation of A' values	*Complex*
SQI	Signal Quality Index	*Real*
V	Mean velocity	*Real*
W	Spectrum Width	*Real*
CCOR	Clutter correction	*Real*
LOG	Signal to noise ratio for thresholding	*Real*
SIG	Signal power of weather	*Real*
C	Clutter power	*Real*
N	Noise power	*Real*
Z	Corrected Reflectivity factor	*Real*
T	UnCorrected Reflectivity factor	*Real*

Figure 5–1: Flow Diagram of RVP7 Processing



5.1 IF Signal Processing

The starting point for all computations within the RVP7 are the instantaneous IF-receiver samples p_n and, the instantaneous burst-pulse or COHO reference samples b_n. These data are available at a very high sampling rate (typically 36MHz), which makes possible the digital implementation of functions that are traditionally performed by discrete components in an analog receiver. The RVP7's all-digital approach replaces a great deal of analog hardware, avoids problems of aging and maintenance, and makes it easy to tune-up the receiver and alter its parameters.

This section describes these IF signal processing steps.

5.1.1 FIR (Matched) Filter

The RVP7 implements a digital version of the "matched" filter that is found in the traditional analog radar receiver. The equivalent Finite-Impulse-Response (FIR) filter is designed using an interactive graphical procedure described in Section 4.5. The filter length (number of taps), center frequency, and bandwidth are all adjustable. The design procedure computes two sets of filter coefficients ${f^i}_n$ and ${f^q}_n$ such that the instantaneous quadrature samples at a given bin are:

$$I = \sum_{n=0}^{N-1} {f^i}_n \times p_n, \qquad Q = \sum_{n=0}^{N-1} {f^q}_n \times p_n$$

where N is the length of the filter. The input samples p_n are centered on the range bin to which the (I, Q) pair is assigned. Note that some of the p_n are likely to overlap among adjacent bins, i.e., the filter length may be chosen to be greater than the bin spacing. Such an overlap introduces a slight correlation between successive bins, but the longer length allows a better filter to be designed.

The convolution sums for I and Q are computed on the RVP7/Main board using dedicated FIR chips that can perform up to 576 million sums of products per second. The p_n are represented as 16-bit signed integers, and the ${f^i}_n$ and ${f^q}_n$ are represented as 10-bit signed integers. A numerical optimization procedure is used to quantize the ideal filter coefficients into their 10-bit hardware values. The overall spectral purity of the FIR filter will typically be greater than 66dBc.

The reference phase for each transmitted pulse is computed using the same two FIR sums, except with b_n substituted for the p_n. For a magnetron system the N b_n samples are centered on the transmitted burst; for a klystron system they are obtained from the CW COHO. If the klystron is phase modulated, then the samples should be from the modulated COHO.

The ${f^i}_n$ coefficients are computed as:

$${f^i}_n \;=\; l_n \times \sin\left[\frac{\pi}{4} + 2\pi\frac{f_{IF}}{f_{SAMP}}\left(n - \frac{N-1}{2}\right)\right] \quad , \quad n = 0 \,...\, N-1$$

where f_{IF} is the radar intermediate frequency, f_{SAMP} is the RVP7/IFD crystal sampling frequency, and l_n are the coefficients of an N-point symmetric low-pass FIR filter that is matched to the bandwidth of the transmitted pulse. The multiplication of the l_n terms by the *sin()* terms effectively converts to the low-pass filter to a band-pass filter centered at the radar IF. The formula for the $f^q{}_n$ coefficients is identical except that *sin()* is replaced with *cos()*.

The phase of the sinusoid terms, and the symmetry of the l_n terms, has been carefully chosen to have a valuable overall symmetry property when *n* is replaced with *(N–1)–n*, i.e., the sequence is reversed:

$$f^i{}_{(N-1)-n} = l_{(N-1)-n} \times \sin\left[\frac{\pi}{4} + 2\pi\frac{f_{IF}}{f_{SAMP}}\left(((N-1)-n) - \frac{N-1}{2}\right)\right]$$

$$f^i{}_{(N-1)-n} = l_n \times \cos\left[\frac{\pi}{4} + 2\pi\frac{f_{IF}}{f_{SAMP}}\left(n - \frac{N-1}{2}\right)\right]$$

$$f^i{}_{(N-1)-n} = f^q{}_n$$

Thus, the coefficients needed to compute *I* are merely the reversal of the coefficients needed to compute *Q*; if you know $f^i{}_n$, then you also know $f^q{}_n$. This is why it is sufficient to print only one set of FIR coefficients during the filter design process described in Section 4.5.

5.1.2 Automatic Frequency Control (AFC)

AFC is used on magnetron systems to tune the STALO to compensate for magnetron frequency drift. The STALO is typically tuned 30 MHz away from the magnetron frequency. The maximum tuning range of the AFC feedback is approximately 7MHz on each side of the center frequency. This assumes that the system's IF frequency is at least 4MHz away from any multiple of half the digital sampling frequency, i.e., 18, 36, 54, or 72MHz.

The RVP7 analyzes the burst pulse samples from each pulse, and produces a running estimate of the power-weighted center frequency of the transmitted waveform. This frequency estimate is the basis of the RVP7's AFC feedback loop, whose purpose is to maintain a fixed intermediate frequency from the radar receiver.

The instantaneous frequency estimate is computed using four autocorrelation lags from each set of N b_n samples. This estimate is valid over the entire Nyquist interval (e.g., 18MHz to 36MHz), but becomes noisy within 10% of each end. Since the span of the burst pulse samples is only approximately a microsecond, several hundred estimates must be averaged together to get an estimate that is accurate to several kiloHertz. Thus, the AFC feedback loop will typically have a time constant of several seconds or more.

Most of the burst pulse analysis routines, including the AFC feedback loop, are inhibited from running immediately after making a pulsewidth change. The center-of-mass calculations are held off according to the value of *Settling time (to 1%) of burst frequency estimator*, and the AFC loop is held off by the *Wait time before applying AFC*. This prevents introducing transients into the burst analysis algorithms each time the pulsewidth changes.

Additional information about using AFC can be found in Sections 2.1.5, 2.2.4, 2.4, and 3.3.6.

5.1.3 Burst Pulse Tracking

The RVP7 has the ability to track the power-weighted center-of-mass of the burst pulse, and to automatically shift the trigger timing so that the pulse remains in the center of the burst analysis window of the **Pb** plot. This means that external sources of drift in the timing of the transmitted pulse (temperature, aging, etc.) will be tracked and nulled out during normal operation; so that fixed targets will remain fixed in range, and clean Tx phase measurements will always be available on every pulse.

The Burst Pulse Tracker feedback loop makes changes to the trigger timing in response to the measured position of the burst. Timing changes will generally be made only when the RVP7 is not actively acquiring data, in the same way that AFC feedback is held off for similar "quiet" times. However, if the center-of-mass has drifted more than 1/3 the width of the burst analysis window, then the timing adjustment will be made right away. Also, there will be an approximately 5ms interruption in the normal trigger sequence whenever any timing changes are made.

The Burst Pulse Tracker and AFC feedback loop are each fine-tuning servos that keep the burst pulse "centered" in time and frequency. These servos have been expanded to include a combined "Hunt Mode" that will track down a missing burst pulse when we are uncertain of both its time and frequency. This coarse-tuning mode is especially valuable for initializing the two fine-tuning servos in radar systems that drift significantly with time and temperature.

When the radar transmitter is *On* but the burst pulse is missing, it may be because either of the following have happened:

- It is misplaced in time, i.e., the Tx pulse is outside of the window displayed in the **Pb** plotting command. In this case, the trigger timing needs to be changed in order to bring the center of the pulse back to the center of the window.
- It is mistuned in frequency, i.e., the AFC feedback is incorrect and has caused the burst frequency to fall outside of the passband of the RVP7 anti-alias filters. In this case the AFC (or DAFC) needs to be changed so that proper tuning is restored.

The Hunt Mode performs a 2-dimensional search in time and frequency to locate the burst; searching across a $\pm$20μsec time window, and across the entire AFC span. If a valid Tx pulse (i.e., meeting the minimum power requirement) can be found anywhere within those intervals then the Burst Pulse Tracker and AFC loops will be initialized with the time and frequency values that were discovered. The fine servos then commence running with a good burst signal starting from those initial points.

Depending on how the hunting process has been configured in the **Mb** menu, the whole procedure may take several seconds to complete. The RVP7's host computer interface remains completely functional during this time, but any acquired data would certainly be questionable. GPARM status bits in word #55 indicate when the hunt procedure is running, and whether it has completed successfully. The BPHUNT (Section 6.27) opcode allows the host computer to initiate Hunt Mode when it knows or can sense that a burst pulse should be present

5.1.4 Interference Filter

The interference filter is an optional processing step that can be applied to the raw *(I,Q)* samples that emerge from the FIR filter chips. The intention of the filter is to remove strong but sporadic interfering signals that are occasionally received from nearby man-made sources. The technique relies on the statistics of such interference being noticeably different from that of weather.

For each range bin at which *(I,Q)* data are available, the interference filter algorithm uses the received power (in deciBels) from the three most recent pulses:

$$P_{n-2},\ P_{n-1},\ \text{and}\ P_n$$

where:

$$P_n \;=\; 10\log_{10}\left(I_n^2 + Q_n^2\right)\ .$$

If the three pulse powers have the property that:

$$|\,P_{n-1} - P_{n-2}\,| \;<\; C_1 \quad \text{and} \quad |\,P_n - P_{n-1}\,| \;>\; C_2 \qquad \text{(Alg.1)}$$

then (I_n, Q_n) is replaced by (I_{n-1}, Q_{n-1}). Here C_1 and C_2 are constants that can be tuned by the user to match the type of interference that is anticipated, and the error rates that can be tolerated. For certain environments it may be the case that good results can be obtained with $C_1 = C_2$; but the RVP7 does not force that restriction.

This 3-pulse algorithm is only intended to remove interference that arrives on isolated pulses, and for which there are at least two clear pulses in between. Interference that tends to arrive in bursts will not be rejected.

Two variations on the fundamental algorithm are also defined. The CFGINTF command (Section 6.23) allows you to choose which of these algorithms to use, and to tune the two threshold constants. You may also do this directly from the **Mp** setup menu (Section 3.3.2).

$$|\,P_{n-1} - P_{n-2}\,| \;<\; C_1 \quad \text{and} \quad P_n - P_{n-1} \;>\; C_2 \qquad \text{(Alg.2)}$$

$$|\,P_{n-1} - P_{n-2}\,| \;<\; C_1 \quad \text{and} \quad P_n - LinAvg(\,P_{n-1},\ P_{n-2}\,) \;>\; C_2 \qquad \text{(Alg.3)}$$

Where *LinAvg()* denotes the deciBel value of the linear average of the two deciBel powers. The Alg.2 and Alg.3 algorithms also include the receiver noise level(s) as part of their decision criteria. Whenever power levels are intercompared in the algorithms, any power that is less than the noise level is first set equal to that noise level. This makes the filters much more robust and properly tunable, so that interference is more successfully rejected on top of blank receiver noise.

Optimum values for C_1 and C_2 will vary from site to site, but some guidance can be obtained using numerical simulations. The results shown below were obtained when the algorithms were applied to realistic weather time series having a spectrum width = 0.1 (Nyquist), SNR = +10dB, and an intermittent additive interference signal that was 16dB stronger than the weather. The interference arrived in isolated single pulses with a probability of 2%.

Performance of the three algorithms is summarized in the first three columns of Table 5–2, for which C_1 and C_2 have the common value shown. The fourth column also uses Algorithm #3, but with the value of C_1 raised by 2dB. The "Missed" rate is defined as the percentage of interference points that manage to get through the filtering process without being removed. The "False" (false alarm) rate is the percentage of non-interference points that are incorrectly modified when they should have been left alone.

Table 5–2: Algorithm Results for +16dB Interference

	Alg.1		Alg.2		Alg.3		Alg.3, C1+=2dB	
C1,C2	Missed	False	Missed	False	Missed	False	Missed	False
6.0dB	17.8%	10.91%	17.8%	4.06%	17.8%	3.48%	10.3%	4.15%
8.0dB	10.5%	6.57%	10.5%	2.42%	10.4%	1.71%	6.1%	1.92%
9.0dB	8.5%	5.09%	8.5%	1.81%	8.3%	1.16%	5.4%	1.28%
10.0dB	7.3%	4.01%	7.3%	1.42%	7.5%	0.79%	5.4%	0.85%
11.0dB	8.9%	3.14%	8.9%	1.06%	8.3%	0.51%	6.5%	0.54%
12.0dB	11.6%	2.53%	11.6%	0.85%	11.3%	0.33%	9.9%	0.35%
13.0dB	17.0%	2.07%	17.0%	0.67%	16.3%	0.22%	15.3%	0.23%
14.0dB	23.5%	1.70%	23.5%	0.54%	22.4%	0.14%	21.6%	0.15%
16.0dB	39.2%	1.21%	39.2%	0.35%	39.6%	0.06%	38.9%	0.06%
20.0dB	67.3%	0.65%	67.3%	0.14%	72.5%	0.01%	72.4%	0.01%

It is important to minimize both types of errors. If too much interference is missed, then the filter is not doing an adequate job of cleaning up the received signal. If the false alarm rate is too high, then background damage is done at all times and the overall signal quality (especially sub-clutter visibility) may be compromised. We suggest that you try to keep the false alarm rate fairly low, perhaps below 1%; and then let the missed percentage fall where it may.

To summarize the numerical results in Table 5–2:

- The "Missed" rates of Alg.1 and Alg.2 are identical, but the "False" rate of Alg.1 is much higher. Alg.1 clearly does not perform as well for additive interference, but it is included in the suite for historical reasons.

- The "Missed" error rate for Alg.3 is nearly identical to that of Alg.2, but Alg.3 has a significantly lower false alarm rate. This is because of the somewhat improved statistics that result when the linear mean of P_{n-2} and P_{n-1} is used in the second comparison, rather than just P_{n-1} by itself. We recommend that Alg.3 generally be chosen in preference to the other two.

- Alg.3 can be further tuned by allowing the two constants to differ. For example, by raising C_1 slightly above C_2 (fourth column), we can trade off a decrease in the "Missed" rate for an increase in the "False" rate. Lowering C_1 would have the opposite effect.

Keep in mind that optimum tuning will depend on the type of interference you are trying to remove. In the previous example, where the interfering signal is only 16dB stronger than the weather, there was a close tradeoff between the "Missed" and "False" error rates. However, Table 5–3 shows the results that would be obtained if the interference dominates by 26db.

Table 5–3: Algorithm Results for +26dB Interference

C1,C2	Alg.1 Missed/False		Alg.2 Missed/False		Alg.3 Missed/False		Alg.3, C2+=5dB Missed/False	
6.0dB	17.8%	10.75%	17.8%	3.95%	17.8%	3.44%	17.8%	0.34%
8.0dB	9.9%	6.48%	9.9%	2.31%	9.9%	1.68%	9.9%	0.15%
9.0dB	7.4%	4.99%	7.4%	1.75%	7.4%	1.14%	7.4%	0.10%
10.0dB	5.9%	3.91%	5.9%	1.36%	5.9%	0.76%	5.9%	0.06%
11.0dB	4.8%	3.06%	4.8%	1.06%	4.8%	0.50%	4.8%	0.04%
12.0dB	3.2%	2.37%	3.2%	0.83%	3.2%	0.33%	3.2%	0.03%
13.0dB	2.6%	1.83%	2.6%	0.62%	2.6%	0.20%	2.8%	0.01%
14.0dB	1.9%	1.45%	1.9%	0.50%	1.9%	0.12%	2.6%	0.01%
16.0dB	1.3%	0.90%	1.3%	0.30%	1.3%	0.05%	5.8%	0.00%
20.0dB	3.1%	0.39%	3.1%	0.12%	2.0%	0.01%	31.5%	0.00%

Notice that we can now re-tune the constants and operate with $C_1 = 13dB$ and $C_2 = 18dB$ (fourth column); which yields a low 2.8% "Missed" rate, and an extremely low 0.01% false alarm rate. Since the false alarm rate is (approximately) independent of the interference power, these filter settings would leave all "clean" weather virtually untouched, i.e., we would have a very safe filter that is intended only to remove fairly strong interference. Such a filter could be left running at all times without too much worry about side effects.

5.1.5 Large-Signal Linearization

The RVP7 is able to recover the signal power of targets that saturate the IF-Input A/D converter by as much as 4–6 deciBels. This is possible because an overdriven IF waveform still spends some of its time in the valid range of the converter, and thus, it is still possible to deduce information about the signal.

Figure 5–2 shows actual signal generator test measurements with normal A/D saturation (lower line), and with the extrapolation algorithms turned on (upper line). The high-end linear range begins to roll off at approximately +10dBm versus +5dBm, and thus has been extended by 5dB.

5.1.6 Correction for Tx Power Fluctuations

The RVP7 can perform pulse-to-pulse amplitude correction of the digital (I,Q) data stream based on the amplitude of the Burst/COHO input. The technique computes a (real valued) correction factor at each pulse by dividing the mean amplitude of the burst by the instantaneous amplitude of the burst. The (I,Q) data for that pulse are then multiplied by this scale factor to obtain corrected time series. The amplitude correction is applied after the Linearized Saturation Headroom correction.

The mean burst amplitude is computed by an exponential average whose *(1/e)* time constant is selected as a number of pulses (See Section 3.3.2). A short time constant will settle faster, but will not be as thorough in removing amplitude variations (since the mean itself will be varying). Longer time constants do a better job, but will require a second or two before valid data is available when the transmitter is first turned on. The default value of 70 will give excellent results in almost all cases.

Whenever the RVP7 enters a new internal processing mode (time series, FFT, PPP, etc.), the burst power estimator is reinitialized from the level of the first pulse encountered, and an additional pipeline delay is introduced to allow the estimator to completely settle. Thus, valid

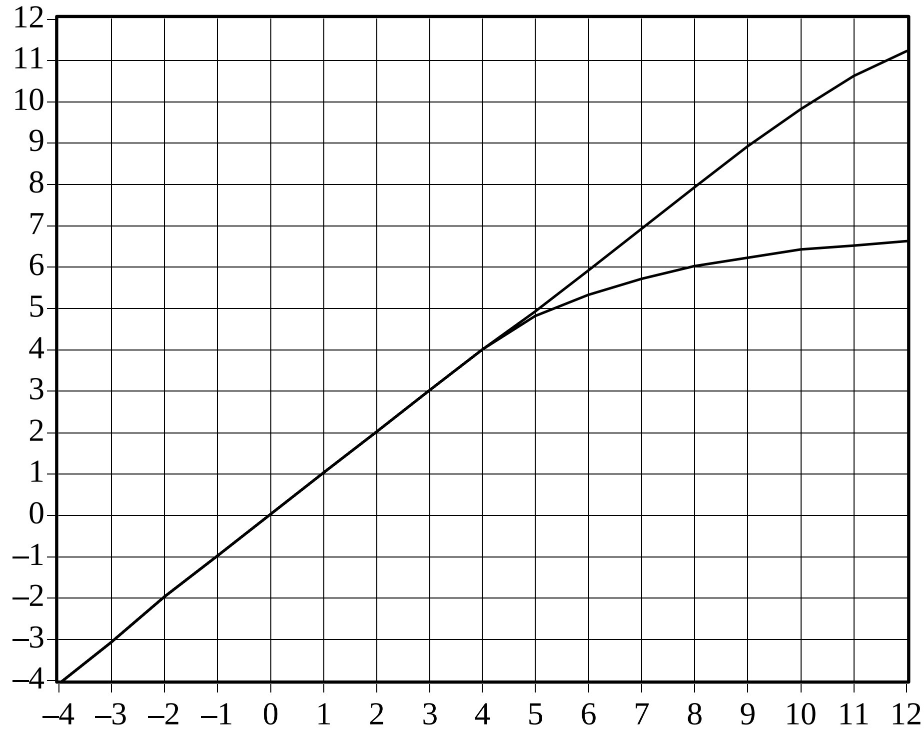


Figure 5–2: Linearization of Saturated Signals Above +4dBm

corrected data are produced even when the RVP7 is alternating rapidly between different data acquisition tasks, e.g., in a multi-function ASCOPE display. The additional pipeline delay will not affect the high-speed performance when the RVP7 runs continuously in any single mode.

For amplitude correction to be applied, the instantaneous Burst/COHO signal level must exceed the minimum valid burst power specified in the "**Mb**" setup section. If that level is not met, e.g., if the transmitter is turned off, then no correction is performed. Thus, the amplitude correction feature conveniently "gets out of the way" when receiver-only tests are being performed.

The maximum correction that will ever be applied is ±5dB. If the burst power in a given pulse is more than 5dB above the mean, or less than 5dB below it, then the correction is clamped at those limits. The power variation of a typical transmitter will easily be contained within this interval (it is typically less than 0.3dB).

Instantaneous amplitude correction is a unique feature of the RVP7 digital receiver. Bench tests with a signal generator reveal that an amplitude modulated waveform having 2.0dB of pulse-to-pulse variation is reduced to less than 0.02dB RMS of (I,Q) variation after applying the amplitude correction.

5.2 Video ("I" and "Q") Signal Processing

This section describes the processing of the video ("I" and "Q") data to obtain the reduced parameters: reflectivity, total power, velocity, width, signal quality index, clutter power correction, and sometimes ZDR. The RVP7 employs two methods (selectable) for processing the I and Q signals: pulse-pair and FFT. The methods are similar except in regard to the procedures for clutter filtering. The pulse pair methods are described below; the FFT clutter filtering algorithms are described in Section 5.8.

5.2.1 Time Series

Recall that the time series synthesized by the FIR filter consist of an array of complex numbers:

$$s_n = [I_n + jQ_n] \quad \text{for} \quad n = 1, 2, 3, \ldots, M$$

where "j" is $-1^{1/2}$. These data samples are analogous to the "I" and "Q" samples in a traditional analog receiver. They are sampled at a selectable resolution in the range 50–133 meters. The time series are the starting point for all calculations performed within the RVP7.

5.2.2 IIR Clutter Filter for PPP-Mode

The RVP7's pulse-pair-processing mode employs a 4th order Infinite Impulse Response (IIR) digital high pass filter to remove low frequency signals due to ground clutter from the time series. Since the width of the clutter can change with the antenna rotation rate, eight different filters (seven high-pass, plus one all-pass) are provided. The filter stop-bandwidths vary from approximately 2% to 14% of the Nyquist interval, and stop band attenuation is at least 40 dB. A setup question allows selection of either 40 dB or 50 dB filters. The 50 dB filters are intended for Klystron systems. Any of the eight filters can be selected independently at each individual range bin. This permits range-dependent clutter removal. The filter algorithm is outlined below.

The input time series s_n is processed to form a filtered output time series s'_n as follows:

$$\begin{aligned} s'_n \;=\; & B_0 s_n + B_1 s_{n-1} + B_2 s_{n-2} + B_3 s_{n-3} + B_4 s_{n-4} \\ & - C_1 s'_{n-1} - C_2 s'_{n-2} - C_3 s'_{n-3} - C_4 s'_{n-4} \end{aligned}$$

where the B's and C's are the filter coefficients. Appendix A gives the magnitude response plots for the set of filters supplied with the RVP7.

5.2.3 Autocorrelations for PPP-Mode

The autocorrelations are computed during pulse-pair-processing mode according to the following algorithms (corresponding physical models are also given):

Parameter and Definition	**Physical Model**
$T_o = \frac{1}{M}\sum_{n=1}^{M} s_n{}^* \; s_n$	$g^r g^t \; (S + C) + N$
$R_o = \frac{1}{M}\sum_{n=1}^{M} s'^*_n \; s'_n$	$g^r g^t \; S + N$
$R_1 = \frac{1}{M-1}\sum_{n=1}^{M-1} s'^*_n \; s'_{n+1}$	$g^r g^t \; S \; e^{j \pi V' - \pi^2 W^2/2}$
$R_2 = \frac{1}{M-2}\sum_{n=1}^{M-2} s'^*_n \; s'_{n+2}$	$g^r g^t \; S \; e^{j 2\pi V' - 2\pi^2 W^2}$

where *M* is the number of pulses in the time average. Here, s' denotes the filtered time series, s denotes the original unfiltered time series and the * denotes a complex conjugate. g^r and g^t represent the transmitter and receiver gains, i.e., their product represents the total system gain. Since the RVP7 is a linear receiver, there is a single gain number that relates the measured autocorrelation magnitude to the absolute received power. However, since many of the algorithms do not require absolute calibration of the power, the gain terms will be ignored in the discussion of these. T_o for the unfiltered time series is proportional to the sum of the meteorological signal S, the clutter power C and the noise power N. R_0 is equal to the sum of the meteorological signal S and noise power N which is measured directly on the RVP7 by periodic noise sampling. T_o and R_0 are used for calculating the dBZ values- the equivalent radar reflectivity factor which is a calibrated measurement. The physical models for R_0, R_1 and R_2 correspond to a Gaussian weather signal and white noise. W is the spectrum width and V' the mean velocity, both for the normalized Nyquist interval [–1 to 1].

The exact value of *M* that is used for each time average will generally be the "Sample Size" that is selected by the SOPRM command (See Section 6.3). However, when the RVP7 is in PPP mode and antenna angle synchronization is enabled, the actual number of pulses used may be limited by the number that fit within each ray's angular limits at the current antenna scan rate. The value of *M* will never be greater than the SOPRM Sample Size, but it may sometimes be less. For example, at 1KHz PRF, 20°/sec scan rate, 1° ray synchronization, and a Sample Size of 80, there will be 50 pulses used for each ray (not 80). Note, however, that the number of pulses used in the "batched" (non-PPP) modes will always be exactly equal to the Sample Size, since those modes are allowed to use overlapping pulses.

5.2.4 Range averaging and Clutter Microsuppression

The next step (optional) is to perform range averaging. Range averaging can be performed over 2, 3, ..., 16 bins. This is accomplished by simply averaging the T_0, R_0, R_1 and R_2 values. This reduces the number of bins in the final output to save processing both in the RVP7 and in the host computer.

At the user's option, the range averaged data can be restricted to include only those bins which have an estimated clutter-to-signal ratio that falls within the CCOR threshold interval. By excluding isolated point clutter targets from the range average the sub-clutter visibility of the averaged data is increased. Specifically, the Doppler test that is applied to each bin in order that it contribute to the overall sum is:

$$10 \log R_0 - 10 \log T_0 > CCOR_{thresh} \ .$$

5.2.5 Reflectivity

The corrected reflectivity Z is output using a log scale based on the following equation:

$$dBZ = 10 \log\left[\frac{T_0 - N}{N}\right] + dBZ_o + 20 \log r + ar + CCOR$$

This equation is simply a dB version of the familiar radar equation for distributed targets. The relationship between the measured autocorrelation function, the received signal and the noise can be expressed as:

$$T_o = g^t g^r S + N$$

where g^t and g^r represent the transmitter and receiver gains, S is the average backscattered power from the targets and N is the measured average noise power. Neglecting attenuation and the contribution of ground clutter (for the moment), the radar equation can be written as.

$$Z = CSr^2 = \left[\frac{Cr_0^2 N}{g^r g^t}\right]\left[\frac{r^2}{r_o^2}\right]\left[\frac{T_o - N}{N}\right]$$

where C is the radar constant and r_o is a reference range which we will later set to 1 km. This is identical to the first three terms of the dB version of the equation with the definition that:

$$Z_o = \frac{Cr_o^2 N}{g^r g^t} = Cr_o^2 I_o \quad where \quad I_o = \frac{N}{g^r g^t}$$

Z_o is called the calibration reflectivity factor. It is the equivalent radar reflectivity factor at the reference range when the return signal power is equal to the noise power (SNR=0 dB). It is sometimes called the minimum detectable dBZ at 1 km. The parameter I_o is the measured noise power at IF with appropriate calibration for the system gain. Calibration of the RVP7 involves defining the radar constant C and measuring the value of I_o. This is discussed in detail in Section 5.4.

Essentially, the measurement of I_o is based on the measurement of the system noise at the time of calibration. However, if the receiver gain were to change after calibration, the use of periodic noise sampling properly corrects for this. For example, if the receiver gain were to change by a factor k, then we would measure a noise value of kN and an autocorrelation value of kT_o, i.e.,

$$Z = CSr^2 = \left[\frac{Cr_0^2N}{g^rg^t}\right]\left[\frac{r^2}{r_o^2}\right]\left[\frac{k\ T_o - k\ N}{k\ N}\right]$$

Thus the k's cancel to give us the same result for Z. This makes the approach robust to system gain fluctuations. Another way of saying this is that as long as the system sensitivity (noise figure) does not change, then the system does not require re-calibration.

The individual terms in the dB form of the equation are summarized below.

1st Term : $10\log\left[\frac{T_0 - N}{N}\right]$: Signal to Noise Ratio

The effect of this term is to subtract the measured noise. It is also used for LOG thresholding. If this number is above the user input value LOG_{thresh} the dBZ is passed.

2rd Term: dBZ_o : Calibration Reflectivity (see discussion above)

dBZ_o is the minimum detectable dBZ at a reference range r_o=1 km,

3th Term: $20\log r$: Range Normalization

This term is the $\left[\frac{r}{r_o}\right]^2$ range normalization expressed in dB form.

4th Term: ar : Gaseous Attenuation Correction

This term accounts for gaseous attenuation. The constant a is set in the RVP7 EEROM since it is a function of wavelength. For a C-band system the default value is 0.016 dB per km (for two-way path attenuation).

5th Term: CCOR: Clutter Correction

This term corrects for the measured ground clutter. It's derivation is discussed in section 5.2.9.

5.2.6 Velocity

For a Doppler power spectrum that is symmetric about its mean velocity, the velocity is obtained directly from the argument of the autocorrelation at the first lag, i.e.,

$$V = \frac{\lambda}{4\pi\tau_s}\theta_1 \quad \text{where} \quad \theta_1 = \arg\{R_1\}.$$

λ is the radar wavelength, τ_s is the sampling time (1/PRF). θ_1 is constrained to be on the interval $[-\pi, \pi]$. When $\theta_1 = \pm\pi$, then $V = \pm V_u$ where the unambiguous velocity is ,

$$V_u = \frac{\lambda}{4\tau_s}.$$

If the absolute value of the true velocity of the scatterers is greater than V_u, then the velocity calculated by the RVP7 is folded into the interval $[-V_u, V_u]$, which is called the Nyquist interval. Folding is usually easily recognized on a color display by a discontinuous jump in velocities. For example, if the true velocity is $V_u + \Delta V$, then the velocity calculated by the RVP7 is $-V_u + \Delta V$, which is $2V_u$ away from the true mean velocity.

For 8-bit outputs, rather than calculating the absolute velocity in scientific units, the RVP7 calculates the mean velocity for the normalized Nyquist interval [–1,1], i.e., the output values are,

$$V' = \frac{\theta_1}{\pi}.$$

For example, an output value of –0.5 corresponds to a mean velocity of $-V_u/2$. The normalized velocity V' is more efficient use of the limited number of bits.

5.2.7 Spectrum Width Algorithms

The spectrum width is a measure of the combined effects of shear and turbulence. To a lesser extent, the antenna rotation rate can also effect the spectrum width. At high elevation angles, the fall speed dispersion of the scatterers also effects spectrum width.

There are two choices for the spectrum width algorithm used in the RVP7, depending on the speed and accuracy that are required for the application:

R_0, R_1 "fast" algorithm valid when SNR >> 10 dB

R_0, R_1, R_2 "accurate" algorithm for SNR >> 0 to 5 dB

The approach used is selected in the SOPRM command. The two approaches are described below:

R_0, R_1 Width Algorithm

Given samples of the Doppler autocorrelation function, numerous estimates of spectral variance can be computed (Passarelli & Siggia, 1983). The particular estimator used by the RVP7 employs the magnitudes of R_0 and R_1 and assumes that the Doppler spectrum is Gaussian (usually an acceptable assumption) and that the signal-to-noise ratio is large. Specifically we have (similar to Srivastava, et al 1979):

$$Variance = 2\ \ln\left[\frac{R_o}{|R_1|}\right] = -2\ \ln[SQI]$$

where "ln" represents the natural logarithm. This can be compared to the expression in the preceding section for SQI to illustrate that this expression for the variance is only valid when:

$$\frac{SNR}{SNR + 1} \approx 1$$

which occurs when the SNR is large.

This variance estimator is normalized to the Nyquist interval in units of $[-\pi, \pi]$. Thus, for example, a variance of $\pi^2/25$ would be obtained from a Gaussian spectrum having a standard deviation equal to one fifth of the total width of the plotted spectral distribution. For scientific purposes, the spectrum width (standard deviation) is more physically meaningful than the variance, since it scales linearly with the severity of wind shear and turbulence. For these reasons, the width W is output by the RVP7:

$$W = \frac{\sqrt{Variance}}{\pi}$$

Again, for efficient packing in 8-bits, width is normalized to the Nyquist interval [–1, 1]. For the example given above, the output width W would be (1/5). To obtain the width in meters per second, one multiplies the output width by V_u .

R_0, R_1, R_2 Width Algorithm

The width algorithm in this case is similar except that the addition of R_2 extends the validity of the width estimates to weak signals. In this case the variance is:

$$Variance = \frac{2}{3} \ln\left[\frac{|R_1|}{|R_2|}\right]$$

The output width W is then defined as in the previous section.

5.2.8 Signal Quality Index (SQI threshold)

An important feature of the RVP7 is its ability to eliminate signals which are either too weak to be useful, or which have widths too large to justify further analysis. This is done via the signal quality index (SQI) which is defined as:

$$SQI = \frac{|R_1|}{R_0}$$

The SQI is the normalized magnitude of the autocorrelation at lag 1 and varies between 0 for an uncorrelated signal (white noise) to 1 for a noise-free zero-width signal (pure tone). Mean velocity estimates are degraded when the spectrum, width is large or when the signal-to-noise ratio is weak. The SQI is a good measure of the uncertainty in the velocity estimates and is a convenient screening parameter to compute. In terms of the Gaussian model, the SQI is :

$$SQI = \frac{SNR}{SNR + 1} \; e^{\frac{-\pi^2 W^2}{2}}$$

where the SNR is the signal-to-noise ratio. For very large SNR's the SQI is a function of the spectrum width only. For a zero-width pure tone (W=0), the SQI is a function of the SNR only (e.g., for W=0, an SNR of 1 corresponds to SQI=0.5). The SQI threshold is typically set to a value of 0.4 to 0.5.

5.2.9 Clutter Correction (CCOR threshold)

In addition to calculating the R_0, R_1 and optional R_2 autocorrelation terms, which are based on filtered time series data, the RVP7 also computes T_0 which is the total unfiltered power. By comparing the total filtered and unfiltered powers at each range bin, a clutter power, and hence a clutter correction, for that bin can be derived. The clutter correction is defined as,

$$CCOR = 10 \log \frac{S}{C + S} = 10 \log \frac{1}{CSR + 1}$$

where S is the weather signal power, C is the clutter power and CSR is the clutter-to-signal ratio. The algorithm for calculating CCOR depends on whether the optional R_2 autocorrelation lag is computed as described below.

R_0, R_1 Clutter Correction

In this case CCOR is estimated from,

$$CCOR_{est} = 10 \log \left[\frac{R_0}{T_0} \right]$$

$$= 10 \log \left[\frac{S + N}{C + S + N} \right] = 10 \log \left[\frac{1 + \frac{1}{SNR}}{CSR + 1 + \frac{1}{SNR}} \right]$$

Here, the expression is strictly valid only when the signal-to-noise ratio (SNR=S/N) is large. Thus when the 2-lag approach is used, the clutter corrections are not as accurate for weak weather signals. However, the error is typically less than 3 dB.

R_0, R_1, R_2 Clutter Correction

In this case there is enough information to compute the clutter signal and noise power independently. The algorithm for CCOR is:

$$CCOR_{est} = 10 \log \frac{S}{C + S} = 10 \log \frac{1}{CSR + 1}$$

The clutter power is computed from:

$$C = T_o - R_o = [C + S + N] - [S + N]$$

The signal power S is then computed from:

$$S = |R_1| \exp^{\frac{\pi^2 W^2}{2}}$$

W is the width that has been previously calculated. This approach yields more accurate results for the clutter correction in the case of a low SNR.

5.2.10 Weather Signal Power (SIG threshold)

A parameter called SIG is also calculated to provide an estimate of the weather signal-to-noise ratio in dB for thresholding. The SIG calculation is different depending on the whether the optional R_2 autocorrelation is computed.

$\boldsymbol{R_0, R_1}$ Calculation

In this case the SIG is computed as follows:

$$SIG = 10 \log \left[\frac{T_0 - N}{N} \right] + CCOR$$

This term represents the SNR after the removal of clutter. The ***CCOR*** value is the one described for $\boldsymbol{R_0, R_1}$ in the previous section.

$\boldsymbol{R_0, R_1, R_2}$ Calculation

In this case the SIG is computed based on the SNR which is:

$$SIG = 10 \log \left[\frac{2\pi S}{R_0 - 2\pi S} \right]$$

where the signal power S is determined as described in the preceding section.

5.2.11 Signal to Noise Ratio (LOG threshold)

A parameter called LOG is also calculated to provide an estimate of the total signal-to-noise ratio in dB useful for reflectivity thresholding. The formula is below:

$$LOG = 10 \log \left[\frac{T_0 - N}{N} \right]$$

5.3 Thresholding

An important feature of the RVP7 is its ability to accept or reject incoming data based on derived properties of the signals themselves. Typically, "rejected" data are not displayed by the user's software, thus making for very clean weather presentations.

5.3.1 Threshold Qualifiers

For data quality control, each RVP7 output parameter can be qualified, i.e., either accepted or rejected for output, based on four threshold criteria:

ID	Criterion Name	Pass Criterion
LOG	Signal-to-Noise Ratio	LOG > threshold
SQI	Signal Quality Index	SQI > threshold
CCOR	Clutter Correction	CCOR > threshold
SIG	Weather Signal Power	SIG > threshold

The calculation of the measured levels (e.g., SQI) for each of these qualifications has been described in previous sections of this chapter. All four qualification criteria can be switched on and off independently, and the threshold levels (e.g., SQI_{thresh}) can each be set independently. Further, each qualifier test can be AND'd and OR'd with any other. This allows very complex thresholds criteria to be constructed as required. The four threshold qualifiers are summarized below.

LOG It is essentially a measure of the total power SNR. This is usually used for thresholding of the reflectivity data. The default LOG threshold value is 0.5 dB.

SQI The SQI threshold is typically used for velocity and width thresholding since it is a measure of the coherency. It is a number between 0 and 1 (dimensionless) where 0 is perfect white noise and 1 is a pure tone (perfect Doppler signal). The default SQI threshold value is 0.5.

CCOR The clutter correction threshold is typically used to reject measurements when the clutter in a range bin is very strong (i.e., when the calculated CCOR is a large negative number in dB). The appropriate value depends on the coherency of the radar system. The default threshold is set to –25 dB. Threshold values less than this (more negative) reject fewer clutter bins. Threshold values closer to zero reject more clutter bins.

SIG This is typically used only for thresholding the spectrum width to assure that the signal power is strong enough for an accurate width measurement. The default threshold value is 10 dB. If R_2 processing is used, this can usually be reduced to 5 dB for width thresholding.

The following are the default threshold combinations for each of the parameters that can be selected for output from the RVP7:

Parameter	Description	Threshold
dBZ	Reflectivity with clutter correction	LOG and CCOR
dBT	Reflectivity without clutter correction	LOG
V	Mean velocity	SQI and CCOR
W	Spectrum width	SQI and CCOR and SIG
ZDR	Differential reflectivity	LOG

5.3.2 Adjusting Threshold Qualifiers

The effect of the various threshold qualifiers for each output parameter are discussed in this section. In optimizing thresholds for your application, it is recommended that you change only one parameter (level or criterion) at a time so that you can verify the effect. Some hints for optimizing the levels for the default criteria are provided below:

LOG To optimize the LOG level, display dBT or dBZ and select the lowest value of the threshold that eliminates the display noise. If the LOG level is set too high you lose sensitivity. Note that if you average more pulses or ranges, then the threshold level can usually be reduced.

SQI To optimize the SQI level, display velocity and select the lowest value of the threshold that eliminates the display noise. If the SQI level is set too high you lose sensitivity. In general, you should see a greater area covered by velocity than reflectivity since the velocity is more sensitive. If you do not, you should reduce your SQI threshold. Note that if you average more pulses or ranges, then the threshold level can usually be reduced.

CCOR This is used to eliminate clutter targets that are very strong. It should not be set to eliminate all clutter targets on a clear day since this means that you are losing sensitivity. To optimize the CCOR threshold it is best to know your system coherency in terms of dB of clutter cancelation. Start at a value of 10 dB greater (closer to 0) than this. Now display a PPI of dBZ at an antenna elevation of ~1 degree. The display should be relatively clean of any clutter targets since most will be rejected. Now reduce the CCOR (more negative) to increase the number of clutter targets on the display until the number of clutter targets does not increase. The optimum value of the CCOR is approximately 5 dB more (closer to zero) than this point. For example, if the number of clutter targets is a maximum at –35 dB, then set the CCOR to ~–30 dB. Note that your clutter filter selection will effect the result.

SIG This should be done last. To optimize the SIG level, display the width W and select the lowest value of the threshold that eliminates the display noise. If the SIG level is set too high you lose sensitivity. Note that if you average more pulses or ranges, then the threshold level can usually be reduced.

When thresholding dBZ and dBT reflectivity data with SQI, the comparison value for accepting those data is the secondary SQI threshold that is defined via a slope and offset from the primary user value (see **Mf** command). This secondary threshold is more permissive (lower valued), and is traditionally used to qualify LOG data only in the Random Phase processing mode. But the secondary SQI threshold is applied uniformly in all processing modes whenever reflectivity data are specified as being thresholded by SQI.

This gives you more freedom in applying an SQI threshold to your LOG data, because the cutoff value for reflectivity can be chosen independently from the cutoff value for the other Doppler parameters. The full SQI test would not normally be applied to LOG data because of the so-called "black hole" problem, i.e., loss of LOG data within regions of high shear, even though the reflectivity itself was strong. You may experiment with applying a secondary SQI threshold to help cleanup the LOG data, but without introducing any significant black holes.

5.3.3 Speckle Filters

A speckle filter is a final pass over each output ray, wherein isolated bins are removed. There are two speckle removers in the RVP7.

- 1D single-ray speckle filter. This can be used for any output parameter.
- 2D 3x3 speckle filter. If enabled, this is applied only to T, V, Z and W.

The 1D speckle filter is the default technique. The 2D 3x3 filter is enabled by selection in the mp TTY setups:

```
2D Final Speckle/Unfold "User" or "Always"
```

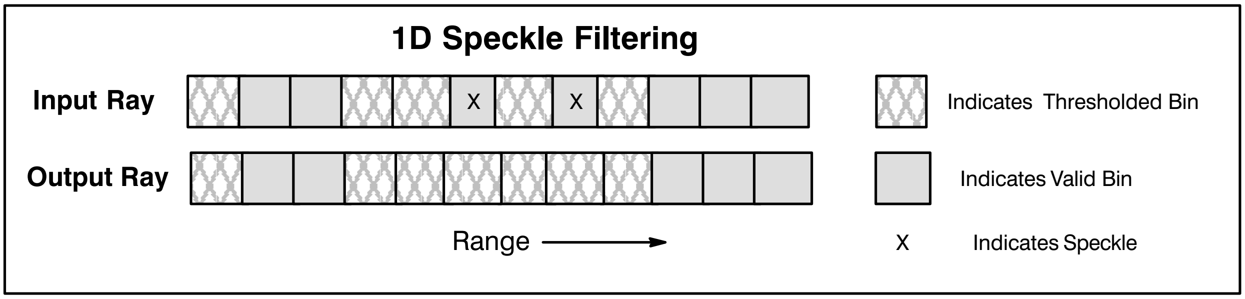
Both of these speckle filters remove isolated data points that are likely to be noise, interference, aircraft, birds or other point targets. Meteorological targets typically occupy multiple range bins so are not effected by the speckle filter. There are two primary benefits derived from using a speckle filter:

- Displays look "cleaner" to observers.
- Thresholds can be set slightly more sensitive without increasing the number of noise pixels.

The 2D 3x3 filter actually performs data filling of "missing speckles" as well as eliminating isolated speckle bins. The two algorithms are discussed below.

1D Speckle Filter

A ray is the basic azimuth unit of the RVP7 (e.g., 1 degree) over which the samples are averaged to obtain the output base data (T, Z, V, W). For this filter, a speckle is defined as any single, valid bin (not thresholded), having thresholded bins on either side of it in range. Any such isolated bin in a ray is set to "threshold". The algorithm is shown schematically below.



Note that there are two independent 1D speckle removers– one for the reflectivity data (dBT, dBZ and ZDR) and one for the Doppler data (V and W). Each one should be switched on or off, depending on the specific nature of the targets being observed. For example, when making a clutter map of the area, one would certainly want to switch both speckle filters off.

2D 3x3 Speckle Filter

The 2D filter examines three adjacent range bins from three successive rays in order to assign a value to the center point. Thus, for each output point, its eight neighboring bins in range and time are available to the filter. Only the dBZ, dBT, Vel , and Width data are candidates for this filtering step; all other parameters are processed using the default 1D speckle filter.

The rules for the filter are as follows:

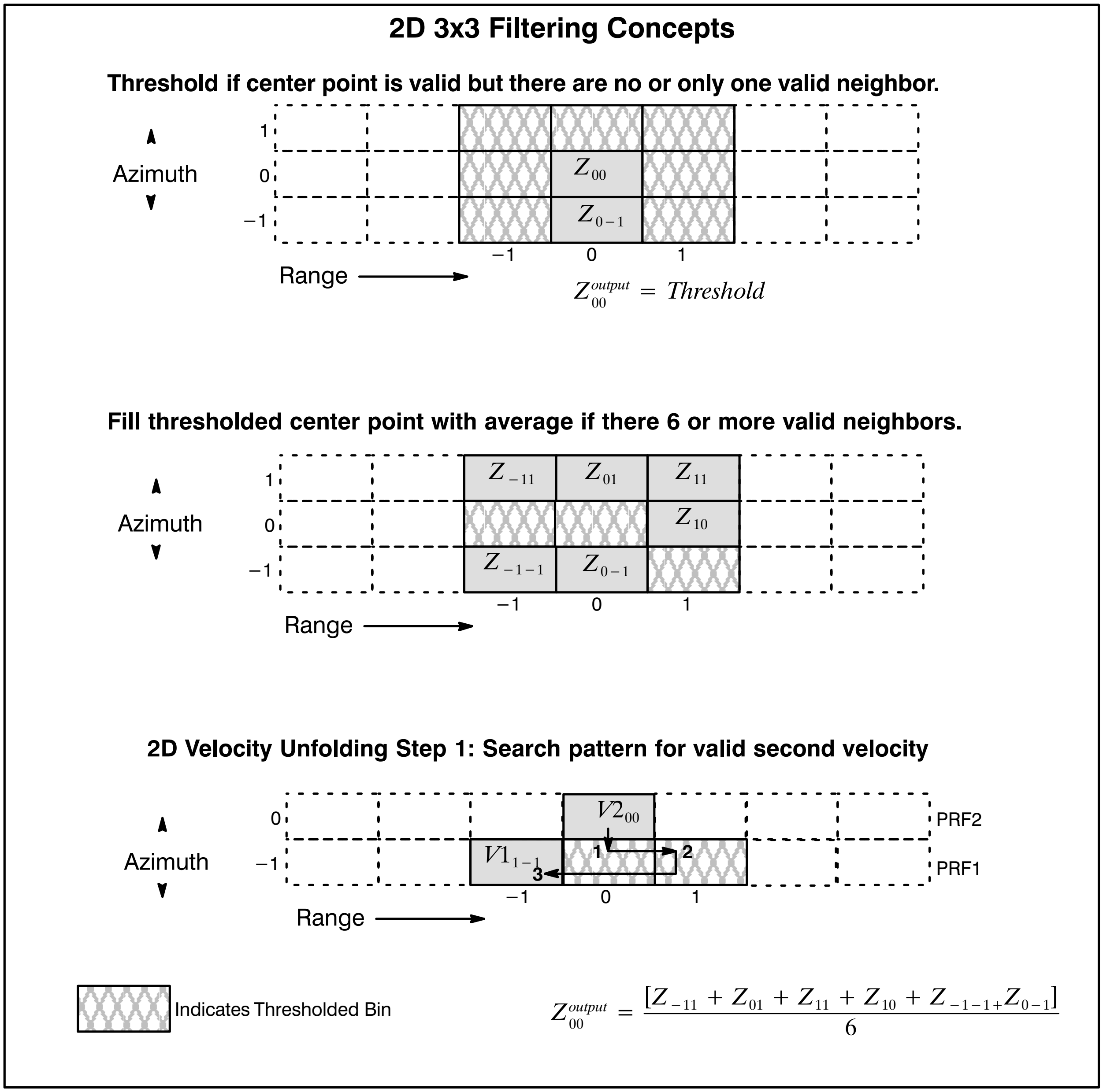
	Center Point Action	
	Assign Threshold	**Else**
Valid Center Point	If there are no or only one other valid point in the 3x3.	Do Nothing. Pass the center point value as-is.
Thresholded Center Point	If there are 5 or fewer valid neighbors in the 3x3.	If there are 6 or more valid neighbors in the 3x3, average to fill the center point.

Thus the 2D 3x3 filter performs 2 functions:

- Filling by interpolation.
- Thresholding of isolated noise bins.

Some examples are shown graphically in the figure below.

For dBZ, dBT, and Width, the interpolated value for filling is computed as the arithmetic average of all available neighbors. For Vel , it is not possible to define a meaningful average in a simple way; so the nearest valid neighbor is simply filled in.



The filter has some interesting properties when combined with other algorithms.

Dual PRF Unfolding

Dual–PRF velocity unfolding is computed within the 3x3 filter whenever both are enabled. There are two steps to the process:

- Step 1: The most recent and the previous ray are used. For every valid point in the most recent ray, the algorithm performs a search among the three nearest neighbors in the previous ray to find a valid velocity. The search pattern is shown at the bottom of the previous figure. This larger selection of alternate–PRF bins makes it more likely that the algorithm will find the pairs of Low/High PRF data that are required for unfolding.

- Step 2: The unfolded velocities are then subjected to the standard 3x3 filtering.

Dual PRF, Random Phase Processing

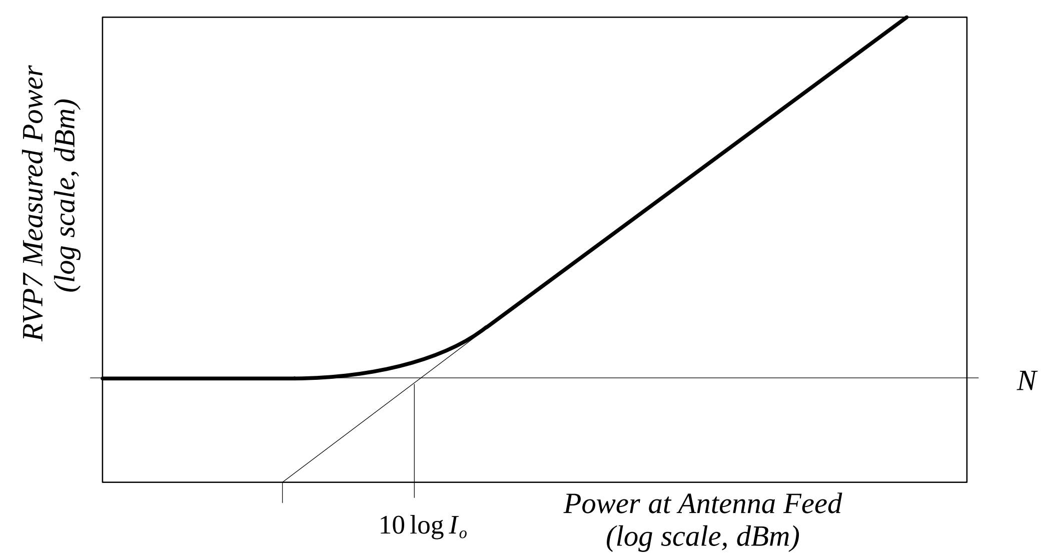
In random phase processing, the "seam" at the start of the second trip is always problematic since the transmitter main bang and nearby clutter will virtually always wipe–out the first few 2nd trip range bins. At a constant PRF the 2nd trip seam is always at the same range, but in dual PRF random phase mode, the seam is different each ray. Thus thresholded bins at the seam of the high PRF can be surrounded on either side by valid bins taken at the low PRF. The 3x3 filter has the effect of interpolating the reflectivity and width data over the bins at the 2nd trip seam. Velocity data will also be filled–in using the nearest neighbor. Thus the 2D filter mitigates much of the damage that is caused at the 2nd trip seam to make a nearly seamless display.

The maximum speed of the RVP7 is reduced to approximately 85000 bins/second when the 3x3 filter is ON– approximately 60% of its maximum throughput when the filter is OFF. This is still a rather large value, and should not affect most customers. For example, there would be no problem running a scan having 2048 bins at 1–degree resolution and a 40 deg/sec scan rate. However, if you really need to operate at the absolute upper limit of the RVP7 s throughput, then the 3x3 filter should be disabled in the Mp menu.

5.4 Reflectivity Calibration

The calculation of reflectivity described in section 5.2.5 required the calibration reflectivity dBZ_o. This section describes it's derivation. Note that customers with the SIGMET IRIS system can use the **zauto** utility to perform the calibration. (See the *IRIS Utilities Manual.)*

Figure 5–3: Model Intensity Curve



Plot Method for Calibration of I_o

This approach generates the curve shown above to determine the value of I_o. The general procedure is to connect a calibrated signal generator to the radar receiver and inject known dBm power levels to generate a calibration plot of measured power vs the inserted power at the antenna feed, similar to that in Figure 5–3. The calibration reflectivity dBZ_o is computed from the radar constant and the value of I_o which is the intercept of the straight line fit with the Noise level. I_o is the signal level for 0 dB SNR, i.e., signal power equals noise power.

Typically a CW test signal is used for this. Follow the instructions provided by the radar manufacturer for injecting a test signal. During calibration, the radar should be fully operational, so that all sources of noise are present. Ideally the transmitter should be turned on during calibration.

Important: Verify with the radar manufacturer that no damage will occur to the signal generator if the transmitter is running during the calibration.

To perform the calibration, insert signals at steps of 5 or 10 dB over the entire range of the system. Draw the plot shown in figure 5–3. You can utilize fine resolution steps at the ends of the scale to observer the details of the roll off. Be sure to raise the antenna up a few degrees to avoid ground thermal noise. Also tune the frequency of the signal generator using the setup command "pr", and displaying the received signal spectrum. Be sure to check the tuning at the end of the calibration to make sure the signal generator and IFD have not drifted apart.

Each time that a new signal level is injected, the measured power values are obtained by first invoking the SNOISE command and then reading-back the results using the GPARM command. The Log of Measured Noise Level (Word 6) from GPARM should be used. This procedure averages many samples together. For IRIS users, this is all handled by the **zauto** utility.

Finally turn it all the way down and make one more sample to measure the noise level N. I_o is obtained from the intercept of the horizontal line at N and the straight line fit to the linear portion of the curve. This value must be corrected for losses as discussed in the section below.

Single-Point Direct Method for Calibration of I_o

This calibration method requires no support software. The approach uses the TTY setups commands. Again the signal generator output must be calibrated in absolute dBm. Use a power meter to check the calibration.

- Turn the radiate off and connect the signal generator to the test signal injection point.
- Raise the antenna to at least 20 degrees, and set the azimuth to point away from any known RF sources including the sun.
- Select the pulse width using the mt command.
- Select the pr command and use the commands to set the following:

```
Plotting Received Power Spectrum...
Rx:Pri,  Zoom:x1–x8,  Navg:25,  Start:100.01 usec (14.99 km),  Span:50 usec
```

- Set the signal generator to the approximate radar RF frequency with a power level corresponding to a strong signal (30 dB above the noise). Use a DC test signal (not pulsed). This signal should be visible as a peak in the spectrum display. Adjust the siggen RF signal frequency so that produces the precise IF frequency (e.g., IF frequency of 30 MHz).
- Turn the signal generator off and record the "Filtered" power level. Note that because of the large averaging it will require several seconds for the average to stabilize.
- Turn the signal generator on, verify that the peak is still at the IF frequency and adjust the power level to obtain precisely 3 dB more "Filtered" power than was observed with the noise only. Again, allow several seconds for the averaging to stabilize after you make each amplitude adjustment.

This is the value of I_o, i.e., the test signal signal power equals the noise power. The next step is to correct the value of I_o for losses as discussed in the section below.

Treatment of Losses in the Calibration

In the calibration of the dBm level of the test signal, be sure to account for any losses that may occur between the antenna feed and the injection point, and in the cable and coupler that is used to connect the signal generator to the injection point. Figure 5–4 illustrates the nomenclature of the various losses that are involved in the calibration. The relationship between the injected test signal and the value of the received power relative to the feed is:

$$dBm_{Feed} = dBm_{Injected} + dBL_{Feed:Coupler}$$

$$dBm_{Feed} = dBm_{Siggen} - dBL_{Coupler} - dBL_{Cable} + dBL_{Feed:Coupler}$$

For example, assume the following:

Loss between the feed and the coupler	$dBL_{Feed:Coupler}$	3 dB
Loss caused by the coupler	$dBL_{Coupler}$	30 dB
Loss in the cable from siggen to coupler	dBL_{Cable}	2 dB

Then if the test signal generator output is –50 dBm, the injected power is

$$dBm_{Injected} = -50-[30+2] = -82 \text{ dBm.}$$

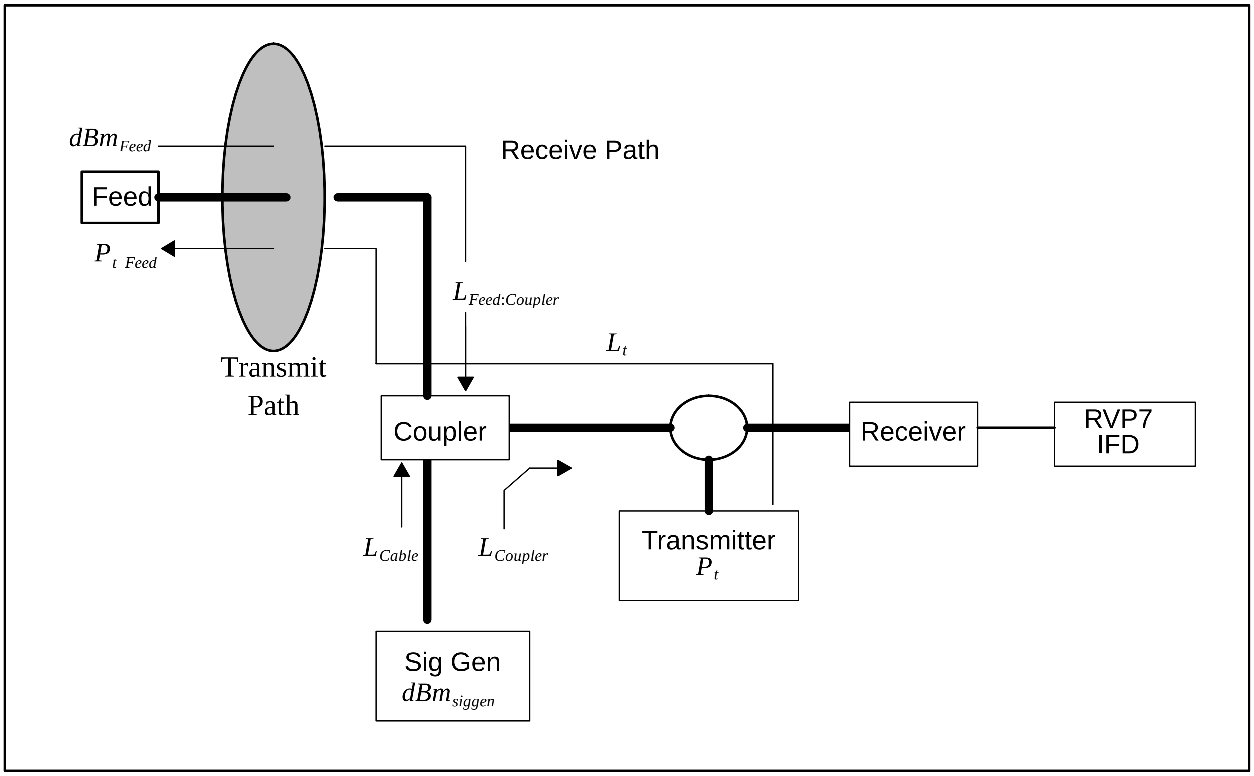
The equivalent power at the feed is then 3 dB more than this

$$dBm_{Feed} = -82+3 = -79 \text{ dBm.}$$

During the calibration, there are several ways to handle the losses using these equations. Two examples are:

- Each signal generator value can be corrected for losses so that the calibration plot shows IFD measured power vs received power at the feed. This is recommended for manual calibration.
- The signal generator values can be plotted directly and the intercept power I_o can be corrected for losses so that it is properly referenced to power at the feed. This is the approach used by the IRIS **zauto** utility.

Figure 5–4: Illustration of Losses that Affect LOG Calibration



Determination of dBZ_o

The calibration reflectivity is determined from the radar equation as follows:

$$dBZ_o \;=\; 10 \;\log\left[Cr_o^2 I_o\right]$$

where I_o is in mW (corrected for receive losses), the reference range r_o is 1 km, and the radar constant C is:

$$C \;=\; \frac{2.69 \;\times 10^{16}\lambda^2}{P_t \;\; \tau \;\theta \;\phi \;G^2} \quad L_t$$

where,

λ	Radar wavelength in cm.
P_t	Transmitted peak power in kW.
L_t	Transmit loss (e.g., 3 dB corresponds to $L_t = 2$)
τ	Pulse width in microseconds.
θ	Horizontal half-power full beamwidth.
ϕ	Vertical half-power full beamwidth.
G	Antenna gain (dimensionless) on beam axis.

The radar constant is determined from the characteristics of your radar (check with the manufacturer if you are unsure of the values). Note that transmit losses are accounted for in the radar constant, while receiver loss is usually included in the calculation of I_o.

Finally, if the value of I_o calculated above was not based on loss-corrected dBm values, correct I_o as follows:

$$dBI_{o\ corrected} = dBI_o - dBL_{Coupler} - dBL_{Cable} + dBL_{Feed:Coupler}$$

Example Calculation of dBZ_o:

This sample calculation is provided so that programmers can check their arithmetic. The radar parameters:

λ	Radar wavelength in cm.	5 cm.
P_t	Transmitted power in kW.	500 kW
L_t	Transmit Loss	2 (3 dB)
τ	Pulse width in microseconds	1 microsecond
θ	Horizontal half-power beamwidth in degrees	1 degree
ϕ	Vertical half-power beamwidth in degrees	1 degree
G	Antenna gain (dimensionless) on beam axis	19,953 (43.0 dB)

The radar constant for this example is,

$$\begin{aligned} C &= \frac{2.69 \times 10^{16}\lambda^2}{P_t\ \tau\ \theta\ \phi\ G^2}\ L_t = \frac{(2.69 \times 10^{16})(5)^2}{(500)(1)(1)(19,953)^2} \quad (2.0) \\ &= 6.76 \times 10^6 \left[mm^6\ m^{-3}\ km^{-2}\ mW^{-1}\right] \end{aligned}$$

Assume that I_o with loss correction is calculated to be –105 dBm (3.16×10^{-11} mW), then dBZ_o is,

$$\begin{aligned} dBZ_o &= 10\log\left[Cr_o^2 I_o\right] = 10\log\left[\ (6.76 \times 10^6)\ (1)^2\ (3.16 \times 10^{-11})\ \right] \\ &= -36.7dB\ (mm^6\ m^{-3}) \end{aligned}$$

This value would be down-loaded to the signal processor using the SOPRM command.

5.5 Dual PRT Processing Mode

The RVP7 supports two major modes for Dual PRT processing, i.e., algorithms using triggers that consist of alternate short and long periods. Most of the Doppler parameters are available in each of these modes. You may also request time series data in both cases; the samples will be organized so that the first pulse of a short PRT pair always comes first.

5.5.1 DPRT-1 Mode

The DPRT-1 trigger consists of a very short PRT from which Doppler data are obtained, followed by a much longer PRT whose purpose is to limit the average duty cycle of the transmitter. No information is extracted from the long PRT pair, but Dual-PRF techniques can still be used by varying the short period from ray to ray. The "-1" suffix in the name for this mode is a reminder that Doppler parameters are computed from the short PRT only. The DPRT-1 mode is intended for millimeter wavelength radars that must run at a very high effective PRF (up to 20KHz) to get an acceptable unambiguous velocity, but which also have a much lower duty cycle constraint on the average number of pulses transmitted each second.

In DPRT-1 mode the requested PRF from the host computer will generally be quite large (up to 20KHz); and the reciprocal of this "effective instantaneous PRF" will determine the trigger's short PRT interval. In this way, all subsequent physical calculations will be scaled correctly, e.g., unambiguous velocity, maximum first trip range, etc., are all supposed to be based on the short PRT interval. The host computer must therefore be configured so that it can ask for these very high trigger rates.

The duration of the long PRT interval is not specified directly by the host computer. Rather, the RVP7's "Maximum number of Pulses/Second" setup parameter is used to compute how much delay to insert in order to insure that the transmitter's duty cycle is not exceeded. This special treatment applies only in DPRT mode; all other modes that have uniform triggers continue to interpret the RVP7's trigger bound as a simple "Maximum PRF".

Since DPRT-1 mode uses only the short pairs of pulses, it is not possible to run the "R2" moment estimation algorithms. The RVP7 will return the GPARM "Invalid Processor Configuration" bit if "R2" is requested in DPRT mode. The error bit will also be returned if the number of pulses requested (sample size) is not even. All other error conditions are the same as FFT mode.

⚠ **Warning: Since the RVP7's "Maximum number of Pulses/Sec" is used to enforce the duty cycle limit, it is essential that it not be overwritten by the host computer's upper PRF limit, which typically will be much higher. To insure this, you must make sure that the PWINFO command is disabled in the RVP7 "Mc" setup menu. You will have no duty cycle protection if you do not do this.**

Note: You may still choose to run Dual-PRF velocity unfolding within the DPRT-1 mode. What will happen is that the short PRT will vary in the selected 3:2, 4:3, or 5:4 ratio, but the overall duty cycle will remain constant. The combination of Dual–PRF and DPRT-1 is tremendously effective in extending the radar's unambiguous velocity interval.

5.5.2 DPRT-2 Mode

The trigger consists of alternating short and long period pulses, where the ratio of the periods is determined by the velocity unfolding ratio that has been selected. Doppler data are extracted from both the short and long pulse pairs (hence the "-2" suffix), and unfolded velocities are made available on each ray based on the combined PRT data from that ray alone. DPRT-2 mode is intended for rapidly scanning radars where the ray-to-ray spatial continuity assumptions of the traditional Dual-PRF algorithms do not apply.

The DPRT-2 velocity unfolding algorithm uses a modified version of the standard Dual-PRF algorithm. Both start by computing a simple velocity difference as a first approximation of the unfolded result. The standard algorithm uses that difference to unfold the velocity from the most recent ray, which yields a lower variance estimate than the difference itself. The DPRT-2 algorithm is similar, except that the folded velocity from both PRTs are unfolded independently and then averaged together.

In addition to the above, the RVP7 also computes the DC average of the (I,Q) data within each bin. This is used as a simple estimate of clutter power, so that corrected reflectivities are available in DPRT-2 mode whenever a non-zero clutter filter is selected. DPRT-1 mode is the same in this respect. However, the DPRT-2 widths use an improved algorithm based on the two different PRTs, and which avoids the SNR sensitivity of the DPRT-1 width estimator.

5.6 Dual PRF Velocity Unfolding

For a radar of wavelength λ operating at a fixed sampling period $\tau_s = 1/PRF$, the unambiguous velocity and range intervals are given by:

$$V_u = \frac{\lambda}{4\tau_s} \quad \text{and} \quad R_u = c\,\frac{\tau_s}{2}$$

where "c" is the speed of light. Often these intervals do not fully cover the span of velocity and range that one would like to measure. The problem is generally worse for short wavelength radars, since that unambiguous velocity span is directly proportional to λ for a given τ_s. If the unambiguous range interval is made sufficiently large by increasing τ_s, then the resulting velocity span may be unacceptably small.

The RVP7 provides a built-in mechanism for extending the unambiguous velocity span by a factor of two, three, or four beyond that given above. The technique, called Dual PRF velocity unfolding, uses two pulse periods rather than one, and relies on the extra information thus obtained to correct (i.e. unfold) the mean velocity measurement from each individual period. The Dual PRF trigger pattern consists of alternating (N+k)-pulse intervals where the period in each interval is either τ_l (for the low-PRF) or τ_h (for the high-PRF). Here "N" is the sample size, and "k" represents a delay that permits the clutter filter to equilibrate to the new PRF after each change. The clutter filter impulse response lengths vary according to which filter is selected.

The two trigger periods τ_l and τ_h must be chosen in either a 3:2, 4:3, or 5:4 ratio. These ratios give factors of two, three, and four times velocity expansion over the τ_h period alone. The unfolding algorithm makes use of the following results. Suppose that the radar observes a target with mean velocity V at each of the two trigger periods. The measured phase angles for the R_1 autocorrelations at the two PRFs are:

$$\theta_l = \frac{4\pi\ V\ \tau_l}{\lambda} \quad \text{and} \quad \theta_h = \frac{4\pi\ V\ \tau_h}{\lambda}$$

where angles outside the basic $[-\pi, \pi]$ interval are returned to that interval by appropriate additions of $\pm 2\pi$. These angles correspond to the ordinary single-PRF Doppler velocity measurements, and the $\pm 2\pi$ uncertainties reflects the fact that each measurement is folded into its own unambiguous interval:

$$V_{ul} = \frac{\lambda}{4\tau_l} \quad \text{and} \quad V_{uh} = \frac{\lambda}{4\tau_h}$$

If we define ϕ to be the difference between the two measured phases then:

$$\phi = \theta_l - \theta_h = \frac{4\pi}{\lambda}\left[\tau_l - \tau_h\right]$$

which can be interpreted as a phase angle within the unfolded interval:

$$V_{u\ unfold} = \frac{\lambda}{4(\tau_l - \tau_h)}$$

Now if τ_l and τ_h are in a 3:2 ratio, then:

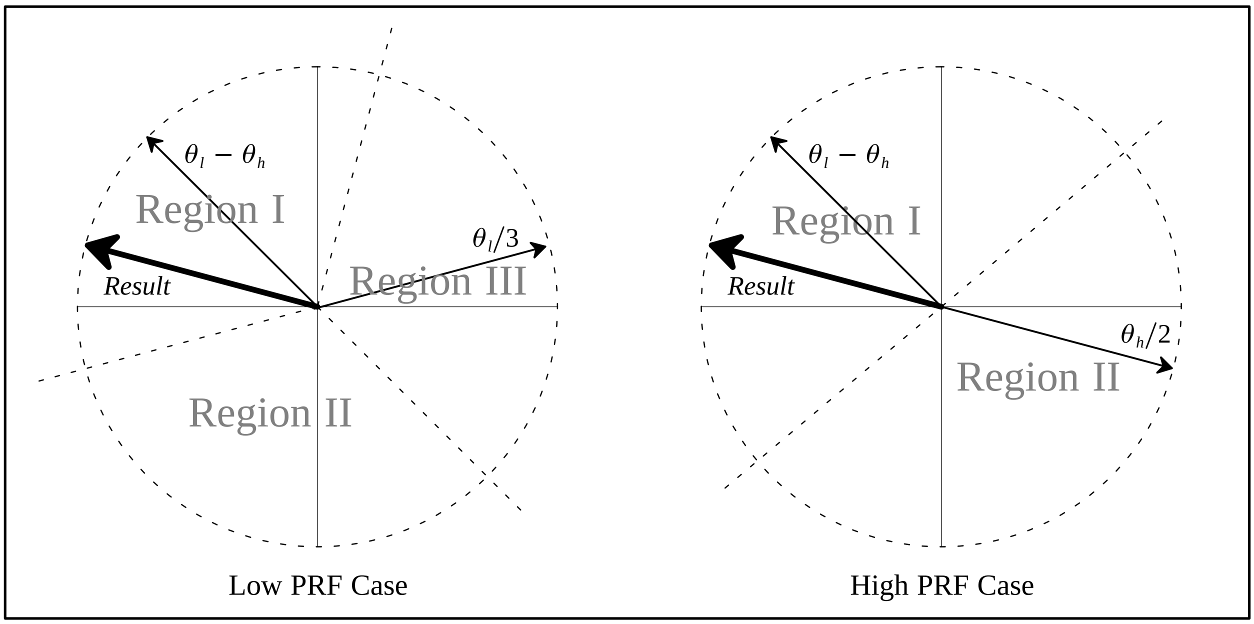
$$\tau_l - \tau_h = \frac{\tau_l}{3} = \frac{\tau_h}{2}$$

$$\text{and thus} \quad V_{u\ unfold} = 3V_{ul} = 2V_{uh}$$

The angle ϕ represents a velocity phase angle in $[-\pi, \pi]$, but with respect to an enlarged unambiguous interval. Thus, by simply differencing the folded angles from the high and low PRFs, we obtain an angle that is unfolded to a larger velocity span. Similar reasoning shows that the 4:3 ratio gives a factor of three improvement over V_{uh}.

In practice, the unfolded angle ϕ is not in itself a suitable velocity estimator. The reason is that the variance of ϕ is equal to the sum of the variances of each of its components, i.e., twice that of the individual measurements alone. If the target is at all noisy, then this increase in variance can be severe. Rather than use ϕ directly, the RVP7 uses it only as a rough estimate in determining how to unfold the individual velocity measured from each PRF.

Figure 5–5: Dual PRF Concepts



This technique is illustrated in Figure 5–5. The figure shows how the low-PRF and high-PRF angles are unfolded based on the difference angle. The diagrams show phase planes representing the large unfolded velocity interval, and the locations of various vectors on those planes. Referring first to the right figure, the difference angle is plotted, and the plane is divided into two equal size regions, one of which is centered on the difference vector. The high-PRF angle is then divided by two and plotted. The resultant unfolded velocity angle must either be this vector, or this vector plus π. Since adding π places the vector into acceptance Region 1 where it is nearest the difference angle, we conclude that this is the correct unfolding. Likewise, on the left diagram we unfold the low-PRF angle by dividing the plane into thirds centered on the difference angle. The result angle is either

$$\frac{\theta_l}{3}\ , \qquad \frac{\theta_l}{3}+\frac{2\pi}{3} \qquad or \qquad \frac{\theta_l}{3}+\frac{4\pi}{3}$$

depending on which one falls into the acceptance Region 1. Note that the resultant angle is the same in each case.

The RVP7 makes efficient use of the incoming data by unfolding velocities from both the low and the high-PRF data, making use each time of information in the previous ray. When low-PRF data are taken the derived velocities are unfolded by combining information from the previous high-PRF interval. Likewise, when high-PRF data are acquired the velocities are unfolded based on the previous low-PRF interval. Thus, when operating in the Dual PRF mode, the RVP7 outputs one data ray for each (N+k)-pulse interval. However, the velocity data in the Dual PRF rays are unfolded, so that the [–1,+1] interval now represents either two or three times the prior velocity range. Put another way, the data are still interpreted as described in the section on mean velocity estimation, except that V_u is now larger.

The width data are also modified somewhat during Dual PRF unfolding. Although valid widths are obtained independently on all rays, those measured at low-PRF are larger than those at high-PRF. This is simply because the dimensionless width units are with respect to a larger velocity interval in the latter case. To compensate for this, low-PRF widths are multiplied by either 2/3 or 3/4 before being output. This puts them in the same scale as the high-PRF values, and thus, the widths do not vary on alternate pulses. A useful consequence of this is that width data can be sent directly to a color display generator without having to plot every other ray in a different scale.

There are a few words of caution that should be kept in mind when using the RVP7 in the Dual PRF processing modes. The unfolding algorithms make the assumption that targets are more-or-less continuous from ray to ray. Otherwise, it would not make sense to use data from a previous ray to unfold velocities in the current ray. Users must therefore assure that their antenna scan rate and beamwidth are such that each target is illuminated, at least partially, over each full 2(N+k)-pulse interval. In practice, a certain amount of decorrelation from ray to ray is acceptable, since the previous rays are used only to decide into which unfolded interval the current ray should be placed. Small errors in the previous ray data, therefore, cause no error in the output. However, large previous-ray errors would lead to incorrect unfolding.

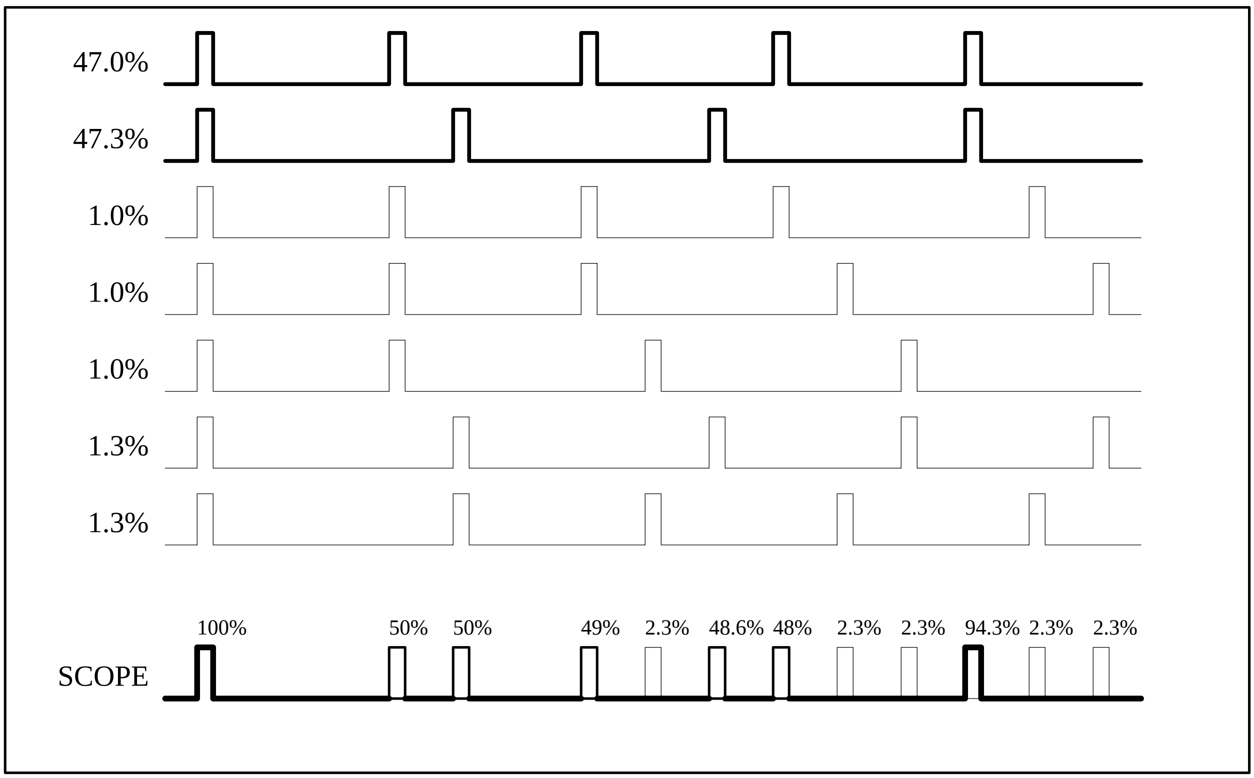
A more subtle side effect of Dual PRF processing arises from clutter filtering because clutter notches now appear at several locations in the unfolded velocity span, rather than just at zero velocity. These additional rejection points come about because the original velocity intervals are mapped some integer number of times to create the unfolded interval. Since each original interval has a clutter notch at DC, it follows that the final expanded velocity interval will have several such notches. For example, in the 3:2 case, in addition to removing DC the clutter filter removes velocities at $-2V_u/3$, $+2V_u/3$, and V_u.

Unfortunately, these clutter filter "images" are a fundamental consequence of the Dual PRF processing technique and are not easily removed. They can cause trouble not only for the velocity unfolding itself, but because the computed clutter corrections to be wrong at the image points. However, there is a useful work-around in the RVP7 to minimize their impact — turning the clutter filter off at far ranges where little clutter is expected and using a narrow clutter filter minimizes the effects of the clutter filter on weather targets.

The 4:3 PRF unfolding ratio is more susceptible to unfolding errors in cases where the spectrum width is large and/or the SNR is low. The user should experiment with the two ratios to determine which provides the best results for their particular application. Although the RVP7 trigger generator can produce any trigger frequency, only the 3:2 and 4:3 ratios can be used with the built-in unfolding algorithms. The RVP7 still permits other PRT ratios to be explored, but the unfolding technique must then be manually programmed on the user's host computer.

Oscilloscope observations of Dual PRF triggers can sometimes be confusing. Figure 5–6 shows seven possible scope traces (and their associated probabilities) for the RVP7 trigger during Dual PRF operation. The PRF ratio is 4:3, and the sample size is 50 pulses at the high PRF, and 37 pulses at the low PRF. The signal labelled "SCOPE" is the composite of these traces, and is what would actually be seen on an oscilloscope. Notice that there are a number of low probability pulses. The exact details of the sample sizes and the trigger hold off time can make the low probability pulses appear to come and go randomly. This is normal, and is no cause for alarm.

Figure 5–6: Example of Dual PRF Trigger Waveforms



5.7 Optional Dual Polarization- ZDR, PHIDP, KDP, LDR, ...

5.7.1 Overview of Dual Polarization

Polarization measurements can provide additional information that can be used to determine more accurate measurements of rainfall or, in some cases, infer particle type such as hail or graupel. The fundamental basis for polarization is that raindrops, particularly larger ones, are not spherical — they are oblate (flattened) such that the horizontal axis is longer than the vertical axis. This means that raindrops will respond differently, for example, to vertical and horizontal polarization of the electric field vector. Because of this, and for technical reasons, most polarization radars use horizontal and vertical polarization. For a review of polarization techniques and variables, please refer to Doviak and Zrnic (1993) section 8.5.

Fundamentally a polarization radar measures amplitude and phase in the same manner as a conventional radar. The new information is that the amplitude and phase can be measured at more than one polarization. The differences in amplitudes and phases measured at different polarizations contain information on the presence or absence of non-spherical scatterers such as large flattened drops. For convenience, some of the basic polarization variables are described below:

ZDR: Differential Reflectivity

In the case of amplitude (power) measurements, the larger horizontal axis of drops causes the power measured at horizontal polarization (of the electric field) to be larger than the power measured at vertical polarization. The ratio of the reflectivity factors Z_H/Z_V expressed in dB is given the name ZDR or differential reflectivity. It is generally positive in rain (i.e., >1) and is usually less than about 5 dB. When the rainfall rate is large, there are typically more large drops so that ZDR is larger. Low ZDR and high dBZ indicates the presence of hail which is perhaps tumbling with no preferred orientation. ZDR, because it is a ratio of powers, is not sensitive to the radar calibration as long as the overall gain of the H and V channels is the same (or calibrated).

PhiDP and KDP: Differential Phase and Specific Differential Phase

In the case of phase measurement, the speed of propagation is also affected by the asymmetry of the larger drops. Because of the longer dimension of the horizontal axis of drops, the medium is effectively more dense for horizontal than for vertical polarization so that the speed of light is reduced for horizontal polarization. This causes the horizontal wavelength to be slightly compressed (more phase cycles per unit distance) in comparison with the vertical wavelength which leads to a phase difference between horizontal and vertical. The phase difference $\Phi_H - \Phi_V$ is called Φ_{DP} differential phase shift. Φ_{DP} increases with range since the phase shifts faster (more frequency cycles per unit distance) for the compressed horizontal microwaves as compared to the faster vertical microwaves. The range derivative of the differential phase, i.e., the change of phase per unit distance, is called K_{DP} or the specific differential phase. K_{DP} is almost directly proportional to the rainfall rate so that it has the potential for improving precipitation rate measurements as compared to traditional Z–R relationship measurements which can be highly inaccurate.

LDR: Linear Depolarization Ratio

Some advanced polarization radars can transmit at one polarization and receive simultaneously in two channels, usually the co–polarized and cross–polarized components. For example, when transmitting horizontal, both horizontal (co–polarized) and vertical (cross–polarized) are received by two separate channels. In the case of vertical or horizontal, the ratio of the power Z_{cross} / Z_{co} is called the linear depolarization ratio or LDR. The amount of incident radiation that is depolarized by a particle depends on the particle shape and orientation (e.g., canting angle with respect to horizontal). Perfectly spherical particles do not depolarize either horizontal or vertical polarization so that LDR is zero. Particles that are wet, tumbling and irregularly shaped will give larger LDR values. Therefore, LDR values in rain tend to be small, e.g., less than –25dB. Larger values of LDR can occur in the bright band or in the presence of hail.

A radar and antenna system must be optimized to measure LDR by assuring that the antenna, feed and supporting struts and radome are not themselves depolarizing the transmitted and received radiation. This is called "cross–pol isolation". The integrated cross–pol isolation of the antenna pattern must be better than about 30 dB for LDR measurement since –20 dB is a large LDR.

[RHOHV, PHIDP] [RHOH, PHIH] [RHOV, PHIV]: Correlation Variables

There are several correlation functions that can be calculated depending on the capabilities of the radar. These are generally complex having both an amplitude and phase. These are all normalized so that a perfect correlation magnitude is 1 and perfectly decorrelated is 0.

RHOHV and PHIDP are the magnitude and phase of the correlation between the horizontal and vertical co-polarized channels. These are available on H/V switching systems or on systems that transmit simultaneous H and V. As discussed in a preceding paragraph, PHIDP can be used to infer precipitation rate. RHOHV in rain is typically very close to 1 (0.98). RHOHV values can be reduced in the case of irregularly shaped, randomly oriented, wet tumbling particles. Thus RHOHV has information on the particle type.

RHOH and PHIH are the magnitude and phase of the correlation between the co-polarized and cross-polar channels for H transmission and simultaneous H and V reception. RHOV and PHIV denote the cross–channel correlation magnitude and phase for vertical transmission. These are available on dual-channel receiver with transmit either fixed or alternating. The information content of the cross-pol correlations is the topic of current research.

5.7.2 Radar System Considerations

A polarization radar is characterized by how it transmits and how it receives. For simplicity we will assume that the radar uses horizontal and/or vertical polarization. However, other polarization pairs could be used (e.g., right and left circular polarization).

Transmit Modes

- **Fixed** (horizontal or vertical)- this can be controlled by a switch or the radar can be simply fixed to transmit a single polarization. If a switch is used, it can be a simple slow waveguide switch rather than a fast switch (pulse-to-pulse).
- **Alternating** (horizontal and vertical)- in this case the radar alternates pulse-to-pulse between horizontal and vertical. A high-power fast switch is used to switch the polarization between the two channels.
- **Simultaneous** (horizontal and vertical)- horizontal and vertical are transmitted simultaneously.

Receive Modes

- **Single-channel receiver**- used only for alternating transmission. The receiver typically receives the co–polarized radiation (transmit H and receive H then transmit V and receive V).
- **Dual-channel receiver**- receives two channels (H and V) simultaneously.

The table below summarizes the various transmit and receive cases and the polarization variables that are available for each. Note that standard parameters are available for all cases (dBT, dBZ, V and W). The RVP7 supports all of these cases.

	Transmitter Type			
Receiver Type	**Fixed H**	**Fixed V**	**Alternating H&V**	**Simultaneous H+V**
Single-Channel	Conventional Radar	Conventional Radar	ZDR RHOHV PHIDP and KDP	Not applicable
Dual-Channel	LDRH RHOH PHIH	LDRV RHOV PHIV	LDRH LDRV RHOH RHOV PHIH PHIV ZDR RHOHV PHIDP and KDP	ZDR RHOHV PHIDP and KDP (*STAR mode*)

The fixed single -channel cases are conventional radars rather than polarization radars. The case of simultaneous H+V transmission and a single radar does not make physical sense. The other cases provide various polarization measurements. The fixed dual-channel cases allow the cross-polarization LDR and the co-pol/cross-pol correlation amplitude and phase to be measured (e.g., RHOH and PHIH). The simultaneous H+V transmission and dual-channel reception is sometimes called the STAR mode (simultaneous transmit and receive). This allows the co–pol measurements to be made (ZDR, RHOHV, PHIDP and KDP). The alternating transmission dual-channel receiver allows both the co–pol and the cross-pol measurements to be made, i.e., it is the most complete.

Summary of Radar System Characteristics

The RVP7 supports all of these modes, but most polarization radar systems do not. As mentioned before, the measurement of cross-pol parameters such as LDR (fixed or alternating transmission and dual-channel reception) requires a radar system that has been optimized for cross–pol isolation, e.g., an offset feed antenna and no radome. By removing the feed, support struts and radome from the path of the radiation, the cross–pol isolation can be improved.

The single-channel alternating method has been used in several polarization radars for ZDR measurement. The advantage of this approach is that it is relatively easy to modify a conventional radar by simply adding a dual port feed and a high-power fast switch above the antenna rotary joints. The disadvantage is that the switch is costly and will eventually fail.

For these reasons, the STAR mode has come into recent use. No switch is required and the components are fairly reliable. The disadvantage of the approach (as it is usually implemented) is that a dual rotary joint and dual waveguides are required to duct both the H and the V through the antenna pedestal up to the antenna feed. In spite of this, the STAR mode offers perhaps the best approach for upgrading an existing radar or for factory installation on a new radar of conventional design.

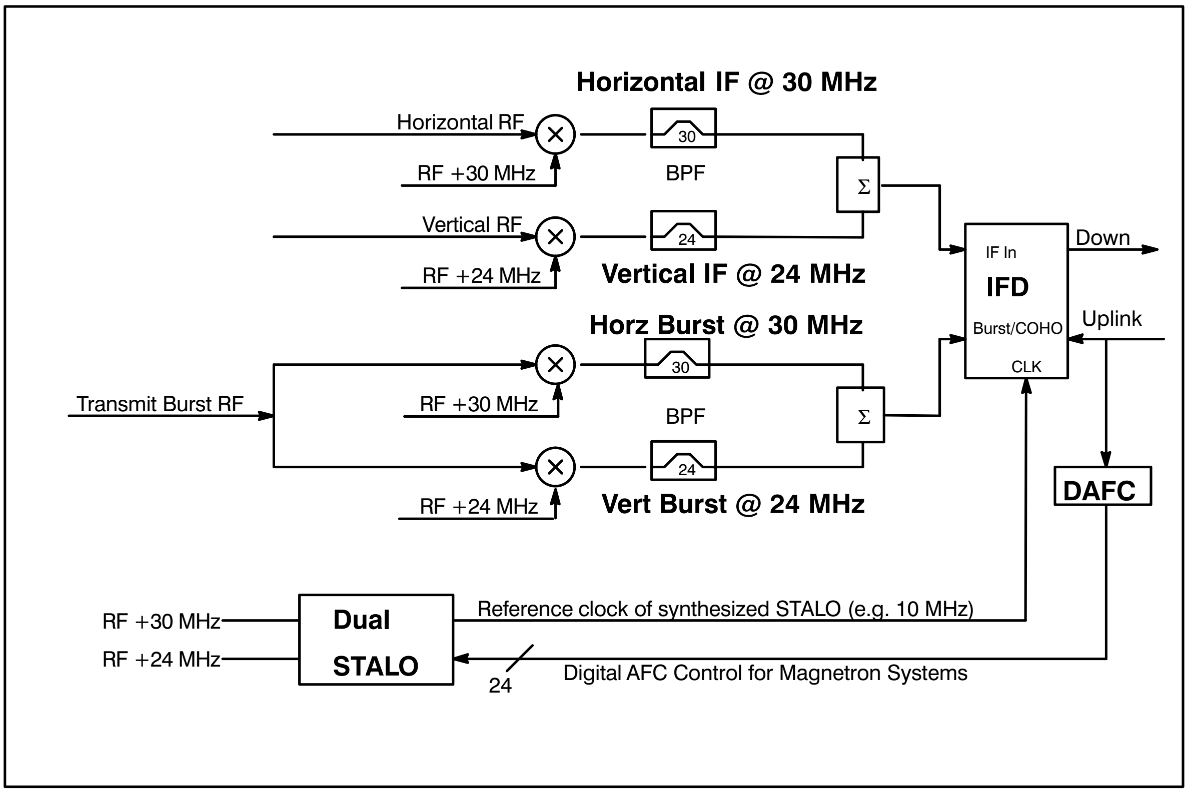
5.7.3 RVP7 Dual-Channel Receiver Approach

Dual-Channel Multiplexing for the IFD

The RVP7 uses an innovative technique for implementing the dual-channel receiver approach, i.e., dual-channel multiplexing. Just as a single wire can carry multiple telephone conversations, two polarization channels can be put on the same wire at different IF frequencies, digitized by the IF Digitizer and then separated by digital filtering. This means that the exact same hardware that is used for a single-channel digital receiver is used for the dual-channel application. The typical IF separation is 6 MHz and the channel isolation is about 50 dB which is more than adequate for even sensitive LDR measurements.

The figure below shows a block diagram of the approach for the magnetron case.

Figure 5–7: Dual Receiver Magnetron Case



Magnetron Approach

The approach for the magnetron case requires a dual output STALO to obtain the two IF's for horizontal and vertical (H and V). The H and V RF channels are mixed to obtain the H and V IF signals at 30 and 24 MHz in the example. In addition, the transmit sample (also know as the "burst pulse" is split and mixed with the two IF's so that there is a version at 30 and 24 MHz. This is important for the determination of the transmit phase corresponding to each IF. Both the H and V IF signals are combined and then digitized in the usual way by a standard RVP7 IFD. The same is done with the 24 and 30 MHz versions of the transmit burst sample. The anti–aliasing filters usually installed on the IFD are removed and replaced by separate filters that are placed ahead of the point where the signals are combined. Note that these filters are centered at the appropriate IF frequency and are typically 6 MHz wide for a 6 MHz IF separation. The two composite signals are then digitized by the IFD identically to the case of a single-channel receiver and later separated during the digital band pass filtering/mixing step to obtain the I and Q of the burst sample and range bin values of I and Q values.

Klystron Approach

In the case of a Klystron system, the approach is the same, except that the COHO must generate two frequencies which are mixed with the STALO to provide the two reference frequencies. These are used in place of the STALO1 and STALO2 frequencies in the diagram. The same two COHO signals (e.g., at 24 and 30 MHz) are then treated identically to the transmit burst pulse in the magnetron case. In this case there is no burst pulse so the two burst pulse mixers are not required.

Reference Clock to IFD

In either case, it is critical that the oscillators (both STALO's in the case of a magnetron and the STALO and both COHO's in the case of a Klystron) be phase locked to a common reference clock. This clock, or a derivative frequency of the clock such as a COHO frequency, is input into the IFD to provide an absolute phase reference. Another alternative is two supply the difference frequency between the two IF's as the reference clock. To do this, the outputs of the two STALO's can be mixed by an additional mixer and filtered to obtain (for example) a 6 MHz reference frequency.

The RVP7 IFD phase locks its sampling crystal to the reference clock input. Trigger generation by the RVP7 will also be phase locked to the reference clock. The reference clock must be in the range 2 to 60 MHz at 0 to –10 dBm and stable in phase to 10^{-7}. The RVP7 IFD must be specially configured with a locking crystal to enable this feature. SIGMET will either factory install the modification or assist the customer in performing the modification and supply the necessary components.

5.7.4 Overview of Processing Algorithms

Polarization Modes and Outputs Supported by RVP7

The RVP7 supports four polarization modes summarized in the table below. For each case, the standard moments (T, Z, V and W) are calculated as well. The notation for the outputs used here is similar to that in standard usage (e.g., Doviak and Zrnic). However, for LDR we use the notation LDRH to indicate that this is the LDR for horizontal transmission. The notation RHOH and PHIH is used to indicate the magnitude and phase of the covariance between the co- and cross–polarized channels for H transmit.

Case	Transmit	Receive	Processing Mode	Polarization Outputs
1	Fixed Horizontal or Fixed Vertical	Dual-Channel	PPP only	LDRH RHOH PHIH or LDRV RHOV PHIV
2	Simultaneous H+V (STAR Mode)	Dual-Channel	PPP or ZDR for FFT, Random Phase and DPRT1&2	ZDR PHIDP KDP RHOHV
3	Alternating H/V	Single-Channel	PPP only	ZDR PHIDP KDP RHOHV
4	Alternating H/V	Dual-Channel	PPP only	LDRH RHOH PHIH LDRV RHOV PHIV ZDR PHIDP KDP RHOHV

Input Receiver Sample Notation

For the discussion of polarization, we will adopt the notation used by Doviak and Zrnic. The received signal for pulse n from a single range bin shall be denoted as:

s^n_{hh}	Receive h: Transmit h	Horizontal co-polar signal
s^n_{vh}	Receive v: Transmit h	Horizontal cross–polar signal
s^n_{vv}	Receive v: Transmit v	Vertical co–polar signal
s^n_{hv}	Receive h: Transmit v	Vertical cross–polar signal

The pulse index is now indicated by the superscript as opposed to the subscript. The first subscript indicates the received polarization while the second subscript indicates the transmit polarization. If the transmit is the same as the received polarization, then this is called the co–polarized signal. If the transmit and receive are different then this is called the cross–polarized signal.

These variables are complex and are the same is the "s_n" notation used earlier, for example we can write:

$$s^n_{hh} = I^n_{hh} + j\ Q^n_{hh}$$

to show the relationship to the received I and Q values. Either filtered and unfiltered versions of the samples can be selected for processing. However, for convenience we will drop the s' notation for filtered samples.

Notation and Model for Correlations

The pulse pair processing mode is used for all of the polarization calculations, except that ZDR-only processing for the STAR case can be done in either FFT or random phase as well as pulse pair. As with the standard moments, the autocorrelations form the basis for the processing of the polarization variables.

The autocorrelations are computed in a manner identical to the standard moments, e.g., in pulse pair mode, the autocorrelations for the horizontal transmit co-polar channel are:

$$T_{o\,hh} = \frac{1}{M} \sum_{n=1}^{M} s_{hh}^{n}{}^{*} \; s_{hh}^{n}$$

$$R_{o\,hh} = \frac{1}{M} \sum_{n=1}^{M} s'^{n}_{hh}{}^{*} \; s'^{n}_{hh}$$

$$R_{1\,hh} = \frac{1}{M-1} \sum_{n=1}^{M-1} s'^{n}_{hh}{}^{*} \; s'^{n+1}_{hh}$$

$$R_{2\,hh} = \frac{1}{M-2} \sum_{n=1}^{M-2} s'^{n}_{hh}{}^{*} \; s'^{n+2}_{hh}$$

What is different is that for polarization systems, this processing can be applied in up to four separate channels (s_{hh}, s_{vh}, s_{vv} and s_{hv}). The physical model for the channel powers is identical to the model used for the standard moment cases, i.e.,

Co-Channel Power	Cross-Channel Power
$R_o^{hh} = g_h^r g_h^t \; S_{hh} + N_h$	$R_o^{vh} = g_v^r g_h^t \; S_{vh} + N_v$
$R_o^{vv} = g_v^r g_v^t \; S_{vv} + N_v$	$R_o^{hv} = g_h^r g_v^t \; S_{hv} + N_h$

Here S is used to denote the actual backscatter average power to the radar which, when multiplied by the appropriate transmitter and receiver gains, yields the actual measured power. Sometimes in comparing powers in two channels (e.g., ZDR and LDR) we will need to know the relative gains of the two channels. However, in many calculations, the relative gains cancel-out and in these cases the algorithms are implemented assuming all the gains are equal to 1.

In the algorithm descriptions, we will often use the notation common in the literature that (for example):

$$R_{o\,hh} = \frac{1}{M} \sum_{n=1}^{M} s'^{n}_{hh}{}^{*} \; s'^{n}_{hh} = < |s'_{hh}|^2 >$$

Noise Bias of Channel Power and Optional Correction

The average noise powers N_v and N_h are assumed to be receiver noise only. These bias the autocorrelations at lag zero, i.e., the channel power measurements. Autocorrelations at lags 1 and 2 are not biased by noise. Cross channel correlations are also not biased by noise, assuming that the noise in the two channels is independent (a good assumption).

The channel noise values are measured directly by the RVP7 during noise sampling. Whether to use these measured values to correct for the noise power when computing a channel power is optionally configured in the TTY setups. The choice is made in the mp TTY setup question "Polarization Parameters NoiseCorrected:YES/NO". If enabled, every time that a channel power is calculated, the noise power is subtracted.

This has some interesting effects. With no noise correction, ZDR values in weak signal regions will be biased by noise toward 0 dB (equal power), while if noise correction is enabled the values will be unbiased but will show substantial deviation over the region. The choice is up to the user.

Clutter Filtering

Clutter filtering is available for all four cases. The use of clutter filters should be carefully considered since many polarization parameters such as ZDR and LDR require highly accurate bin–to–bin consistency. The clutter filters will attenuate some small amount of weather near zero velocity and the amount could be slightly different at the two polarizations. When using the clutter filters, users should verify that the functioning is acceptable for their application.

The standard moments (T, Z, V, W) are filtered in the usual manner by selecting a clutter filter (other than Filter #0 which is the All Pass filter). To enable clutter filtering of the polarization variables, the mp TTY setup question "Polarization Parameters Filtered" should be set to YES. The polarization algorithms will then be calculated with filtered time series. Note that filtering can be effectively disabled for the polarization variables by selecting Filter #0 even if the mp setup is set to enable filtering. This allows users to make easy comparison of the filtered vs unfiltered results using their application software without having to change the RVP7 setup.

The s' notation for filtered samples shall be dropped in the algorithm discussions. It is understood that the input samples in all cases may be either filtered or unfiltered according to the user's choice in the TTY setups.

5.7.5 Case 1: Fixed Transmit: Dual-Channel Receiver

Input Receiver Samples

In fixed mode the radar is configured (either permanently or by means of a switch) to transmit either vertical or horizontal polarization with dual-channel reception of both the co- and cross–channel polarizations, e.g., transmit horizontal and receive both horizontal (co) and vertical (cross) polarizations.

The received samples in the two transmit cases are:

Transmit Horizontal or **Transmit Vertical**

$$\left[s_{hh}^{1} : s_{vh}^{1}\right] \left[s_{hh}^{2} : s_{vh}^{2}\right] \left[s_{hh}^{3} : s_{vh}^{3}\right] \;.\;.\;.\; \left[s_{hh}^{M} : s_{vh}^{M}\right] \quad \text{or} \quad \left[s_{vv}^{1} : s_{hv}^{1}\right] \left[s_{vv}^{2} : s_{hv}^{2}\right] \left[s_{vv}^{3} : s_{hv}^{3}\right] \;.\;.\;.\; \left[s_{vv}^{M} : s_{hv}^{M}\right]$$

Calculation of the Polarization Measurands

The processing in this mode is done by pulse pair algorithm. The user may select a clutter filter, but in general this is not recommended for polarization studies since the clutter filter might interfere with the accuracy of sensitive parameters such as LDR.

The polarization measurands for the two transmit cases are as follows:

Transmit Horizontal or **Transmit Vertical**

$$LDRH = 10\ LOG\left[\frac{S_{vh}}{S_{hh}}\right] \quad \text{or} \quad LDRV = 10\ LOG\left[\frac{S_{hv}}{S_{vv}}\right]$$

$$= 10\ LOG\left[\frac{<\ |s_{vh}|^2\ >\ -\ N_v}{<\ |s_{hh}|^2\ >\ -\ N_h}\right] - XDR \quad \text{or} \quad = 10\ LOG\left[\frac{<\ |s_{hv}|^2\ >\ -\ N_h}{<\ |s_{vv}|^2\ >\ -\ N_v}\right] + XDR$$

$$RHOH = |\rho_h| \quad \text{or} \quad RHOV = |\rho_v|$$

$$PHIH = \arg[\rho_h] \quad \text{or} \quad PHIV = \arg[\rho_v]$$

Here, the H and V average channel powers are computed as follows with optional noise correction, i.e.,

Co- $\quad g_h^r g_h^t\ S_{hh} = <\ |s_{hh}|^2\ >\ -\ N_h \quad$ or $\quad g_v^r g_v^t\ S_{vv} = <\ |s_{vv}|^2\ >\ -\ N_v$

Cross- $\quad g_v^r g_h^t\ S_{vh} = <\ |s_{vh}|^2\ >\ -\ N_v \quad$ or $\quad g_h^r g_v^t\ S_{hv} = <\ |s_{hv}|^2\ >\ -\ N_h$

The complex covariance ρ (used above) is:

for H transmit $\quad \rho_h = \dfrac{<\ s_{vh}\ s_{hh}^{*}\ >}{\sqrt{S_{vh}S_{hh}}} \quad$ or for V transmit $\quad \rho_v = \dfrac{<\ s_{hv}\ s_{vv}^{*}\ >}{\sqrt{S_{hv}S_{vv}}}$

Fortunately, the algorithms do not require us to know all of the individual gain terms. They cancel in the calculation of ρ so are taken as =1 in the implementation. However, the differential receiver gain XDR must be known from calibration to calculate LDR:

$$dB\ Value\ is\quad XDR = 10\ LOG\ xdr \quad where\ the\ linear\ value\ is \quad xdr = \frac{g_v^r}{g_h^r}$$

5.7.6 Case 2: Simultaneous Dual Transmit and Receive (STAR mode)

Input Receiver Samples

In this mode there is simultaneous transmit and receive of both vertical and horizontal polarization. For each pulse there is a measurement of the complex amplitude in each channel, i.e.,

$$\left[s_{hh}^{1} : s_{vv}^{1}\right] \left[s_{hh}^{2} : s_{vv}^{2}\right] \left[s_{hh}^{3} : s_{vv}^{3}\right] \ . \ . \ . \ \left[s_{hh}^{M} : s_{vv}^{M}\right]$$

We will assume that M samples are collected for processing, i.e., Note that even though there is cross–polarized radiation received in each channel, this cross–polar contribution can be neglected since the co-polarized received signal is much stronger.

Calculation of the Polarization Measurands

The processing in this case is done by pulse pair mode. However both FFT and random phase processing can be performed if only ZDR and standard moments are requested for ouput. In any mode, the user may select a clutter filter, but in general this is not recommended for polarization measurements since the clutter filter might interfere with the accuracy of sensitive parameters such as ZDR.

The RVP7 calculates the following polarization parameters:

$$ZDR = 10 \ LOG \left[\frac{S_{hh}}{S_{vv}}\right]$$

$$ZDR = 10 \ LOG \left[\frac{< |s_{hh}|^2 > - N_h}{< |s_{vv}|^2 > - N_v}\right] + GDR$$

$$RHOHV = |\rho_{hv}(0)|$$

$$PHIDP = \arg[\rho_{hv}(0)]$$

KDP based on least squares fit to PHIDP (see Section 5.7.10).

where the following definitions are used:

$$g_h^r g_h^t \ S_{hh} = < |s_{hh}|^2 > - N_h \qquad g_v^r g_v^t \ S_{vv} = < |s_{vv}|^2 > - N_v$$

The noise powers the two channels are denoted as N_h and N_v . The noise corrections to S_{hh} and S_{vv} are optionally configured in the TTY setups. GDR is the total (transmit and receive) differential channel gain. It must be calibrated for the system.

$$dB \ Value \ is \quad GDR = 10 \ LOG \ xdr \quad where \ the \ linear \ value \ is \quad gdr = \frac{g_v^r g_v^t}{g_h^r g_h^t}$$

The correlation function is computed from:

$$\rho_{hv}(0) = \frac{< s_{vv} \ s_{hh}^* >}{\sqrt{S_{hh} S_{vv}}}$$

The gain terms cancel in the calculation of ρ so in the implementation they are simply assumed to be =1.

5.7.7 Case 3: Alternating H/V Transmit: Single-Channel Receiver

Input Receiver Samples

This is the traditional ZDR radar with a high-power fast switch that alternates between horizontal and vertical on each pulse. The switch is made just prior to the transmit pulse so that the transmitter radiates and then receives at a single polarization for each pulse. Thus the samples are:

$$s_{hh}^{1} \qquad s_{vv}^{2} \qquad s_{hh}^{3} \quad \ldots \quad s_{vv}^{M+1}$$

For the discussion below we will assume that there are M+1 total samples with M/2 horizontal pulses indexed by (1, 2, 3...M–1) and M/2+1 vertical pulses indexed at (2, 4, 6, ...M). Note that the processor does not assume that the first pulse in a sequence is horizontal.

Calculation of the Polarization Measurands

The processing is done in pulse pair with optional clutter filter. Again, for accurate ZDR measurements, the clutter filter may interfere.

The RVP7 calculates the following:

$$ZDR = 10\ LOG\left[\frac{S_{hh}}{S_{vv}}\right]$$

$$ZDR = 10\ LOG\left[\frac{<|s_{hh}|^2> - N_h}{<|s_{vv}|^2> - N_v}\right] + GDR$$

$$PHIDP = \frac{1}{2}\ \arg\left[R_a\ R_b^*\right]$$

$$RHOHV = \frac{|\rho_{hv}(T_s)|}{\left[\rho_{hv}(2T_s)\right]^{0.25}}$$

KDP based on least squares fit to PHIDP (see Section 5.7.10).

where the following definitions are used:

$$|\rho_{hv}(T_s)| = \frac{|R_a| + |R_b|}{2\sqrt{S_{hh}S_{vv}}}$$

$$\rho(2T_s) = \frac{\left|\sum_{n=1}^{M/2-1}(s_{hh}^*[2n-1]\ s_{hh}[2n+1] + s_{vv}^*[2n]\ s_{vv}[2n+2])\right|}{(M/2-1)(S_{hh}+S_{vv})}$$

$$R_a = \frac{1}{M/2}\sum_{n=1}^{M/2} s_{hh}^{2n-1}*\ s_{vv}^{2n} \qquad \text{and} \qquad R_b = \frac{1}{M/2}\sum_{n=1}^{M/2} s_{vv}^{2n}*\ s_{hh}^{2n+1}$$

The calculation of the channel powers ($<|s_{hh}|^2>$ and $<|s_{vv}|^2>$) is done using alternating pulses in this case. Note that in the calculation of R_b, the RVP7 uses the extra M+1 sample. The gain terms cancel in the calculation of ρ so in the implementation they are simply assumed to be =1.

5.7.8 Case 4: Alternating H/V Transmit: Dual-Channel Receiver

Input Receiver Samples

This is the most comprehensive case of polarization operation since it permits calculation of all of the polarization measurands. In this case the transmitter alternates pulse-to-pulse between horizontal and vertical polarization and the dual-channel receiver provides measurement of both the co- and the cross-polarized return, i.e.,

$$\left[s_{hh}^{1} : s_{vh}^{1}\right] \left[s_{vv}^{2} : s_{hv}^{2}\right] \left[s_{hh}^{3} : s_{vh}^{3}\right] \left[s_{vv}^{3} : s_{hv}^{3}\right] \ . \ . \ . \ \left[s_{vv}^{M+1} : s_{hv}^{M+1}\right]$$

We will assume that M+1 samples are collected for processing (an extra sample is required for the calculation Rb per section 5.7.7).

Calculation of the Polarization Measurands

The RVP7 calculates the following:

Co-polar channel measurements

ZDR, PHIDP, RHOHV Identical to alternating case Section 5.7.7.

Cross-polar channel measurements

LDRH, LDRV, RHOH, RHOV, PHIH, PHIV Identical to fixed case Section 5.7.9

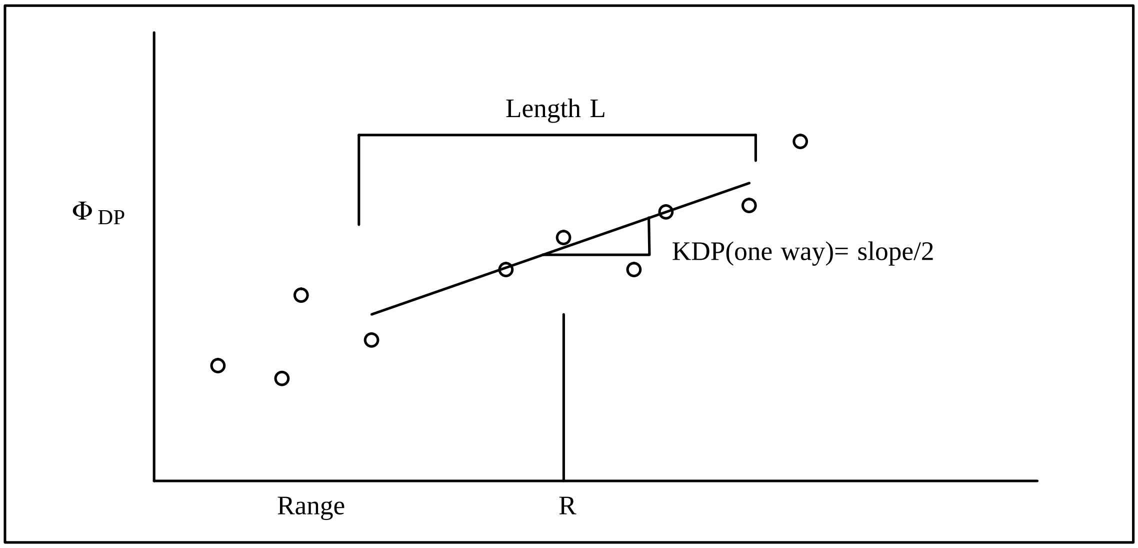
The co-polar channel measurements are exactly as they are for the alternating single-receiver case. The cross–polar measurements are calculated using fixed case algorithms except they are calculated for BOTH H and V polarizations.

5.7.10 KDP Calculation

In all modes that compute PHIDP, the signal processor can also be configured to compute KDP- the specific differential phase in units of degrees per km. This is the range derivative of PHIDP. There are two techniques that have been used to obtain this:

- The smoothed range derivative.
- The slope from a least squares fit.

The RVP7 uses the least squares approach which is shown schematically in the figure below.



The graph shows the thresholded differential phase vs range. This is the starting point for the algorithm. The length scale L is selectable by the user in the TTY setups (mp section, KDP Length in km, default 5.00 km). The KDP value for a bin at range R is computed from a least squares fit that includes points that are within +–L/2 as indicated in the figure. PHIDP is output by the processor on the unambiguous interval of 0 to 180 degrees. Before fitting, the points are first unfolded to a common interval by starting at the left-most point and then moving right assuming that a difference of more than 1/2 the unambiguous interval is the result of folding. Since it is the slope that is of interest, the absolute interval is not critical, as long as the points are in a common interval.

After fitting, the slope is obtained which corresponds to the 2-way KDP since it is based on the 2-way measurement of PHIDP . To be consistent with most values in the literature, the slope value is divided by 2 so that the final output is the one-way KDP in degrees per km (with a wavelength scaling in the data format).

This procedure is repeated for each bin. Thus if the bin spacing is 250 m, the output bin spacing of KDP will be 250 m. It is required that there be at least 50% of the possible number of bins present in the interval L to calculate a valid KDP, else the KDP is set to the threshold value. Since the input PHIDP values are already thresholded, the only additional threshold on KDP is this 50% rule.

5.7.11 Standard Moment Calculations (T, Z, V, W)

Overview

Standard moments are available for all four of the polarization cases. Since there can be up to four different channels of time series input, there are several choices for computing the standard moments. For example, in the STAR mode (Case 2), the standard moments can be computed from:

- s_{hh} samples
- s_{vv} samples
- Average of the results from the s_{hh} and s_{vv} samples

The third case is handled by averaging the individual channel correlations, and then using the average correlations in the standard moment processing. The averaging must take into account the differential gain of the channels.

The selection of which method to use is made in setup. There are four questions posed in the mp section:

```
T/Z/V/W computed from:   H–Xmt:YES   V–Xmt:YES
T/Z/V/W computed from:   Co–Rcv:YES  Cx–Rcv:NO
```

The first two questions are used to specify that ***given a choice*** between vertical and horizontal transmit, which transmit polarization to use. Thus for the fixed H or V case where there is only one transmit polarization, this question does not apply. The processor will simply use samples for the polarization is transmitted.

The second two questions are used to specify that ***given a choice*** between using the co- or cross-polar receivers which one shall be used. This question applies only to systems that can measure LDR, i.e., fixed or alternating transmit, dual-channel receiver systems).

The tables in the sections below summarize the standard moment calculations for each of the four modes and how to configure the four TTY setup responses. Note that these are the only supported modes. Some combinations of responses are unsupported. For example, it is not supported to answer both Co-Rcv: NO and Cx-Rcv: NO.

The top of each table identifies the transmitter/receiver case and what samples are available. The notation HH signifies that the s_{hh} samples are available. The tables use "——" to indicate that either a YES or NO response will cause the same result, i.e., the RVP does not care what response is made. In cases where averaging is performed, the type of weighting used is indicated (either GDR or XDR weighting).

Model for standard moment autocorrelations

The model for the moment autocorrelation calculations is as follows (using R_o as an example):

$$R_0^{hh} = g_h^r\, g_h^t\, S_{hh} + \overline{N}_h \qquad R_0^{vh} = g_v^r\, g_h^t\, S_{vh} + \overline{N}_v$$

$$R_0^{vv} = g_v^r\, g_v^t\, S_{vv} + \overline{N}_v \qquad R_0^{hv} = g_h^r\, g_v^t\, S_{hv} + \overline{N}_h$$

where:

$R_0^{hh},\ R_0^{vh},\ R_0^{vv},\ R_0^{hv}$	Are the autocorrelations if the samples at lag zero.
$S_{hh},\ S_{vh},\ S_{vv},\ S_{hv}$	The average power returned from the scatterers.
$g_h^r,\ g_v^r$	Receiver gains for horizontal and vertical receive.
$g_h^t,\ g_v^t$	Transmitter gains for horizontal and vertical transmit.
$\overline{N}_h,\ \overline{N}_v$	Measured noise power of the samples.

In other words, the power that is measured in a channel has two components:

- Backscattered power from the targets that is effected by the transmitter and receiver channel gains.
- Receiver noise which is measured by the RVP7 during noise sampling.

In the case of R1 and R2 autocorrelations, the model is similar except that there is no noise bias.

Calibration Parameters

For dBZ calculations, a calibration constant is required, i.e., the dBZ_o value in Section 5.4. Depending on the polarization case and the technique selected for standard moment calculation, it may also be required to have GDR and XDR, i.e.,

- GDR- The ratio of the total gains (transmit/receive) of the two co–receive channels.
- XDR- The ratio of the receiver gains in a dual receiver system. This is not required for the Case 2: STAR or the Case 3: Alternating Single–Channel.

The RVP7 supports a single calibration reflectivity dBZ_o. In all cases it is assumed that the dBZ_o is for the horizontal co–receive (HH) channel. The only exception is for fixed vertical polarization, in which the algorithm assumes that the calibration is for the vertical co-receive (VV) channel. XDR and GDR are also downloaded and used to adjust the dBZ_o as required depending on the user's selection for the standard moments. For example, in STAR mode, if the user selects dBZ to be computed from the VV channel, the dBZ_o for the HH and a GDR adjustment are used to calculate the dBZ in the VV channel.

The remainder of this section discusses the standard moment calculation options for the each polarization case. For a discussion of how to calibrate XDR and GDR see Section 5.7.13.

Case 1H: Fixed Horizontal Transmit, Dual Channel Receive- (HH, VH)				
dBZo from HH Channel	**TTY Setup Question Responses**			
Calculate T, Z, V, W from:	**H–Xmt**	**V–Xmt**	**Co–Rcv**	**Cx–Rcv**
HH (co) (Recommended)	—	—	YES	NO
VH (xdr^{-1} weighting)	—	—	NO	YES
HH+VH (xdr^{-1} weighting)	—	—	YES	YES

HH Channel (co–pol)

This is the recommended channel for the case of linear polarization. The reason is that for linear polarization, the co–polar channel will have the strongest signal. Processing is identical to a conventional radar.

VH Channel (cross–pol)

This choice would be used for circular or elliptic transmit polarization. Since the algorithm assumes that dBZo is from the co–polar channel, xdr is used to adjust the autocorrelations as follows:

$$T_0 = xdr^{-1}\ T_o^{vh}$$

$$R_0 = xdr^{-1}\ R_o^{vh}$$

$$R_1 = xdr^{-1}\ R_1^{vh}$$

$$R_2 = xdr^{-1}\ R_2^{vh}$$

$$N = xdr^{-1}\ N_v$$

These adjusted autocorrelations are then used as per the standard moment processing for a conventional radar. To illustrate this, consider the example of reflectivity processing. The radar equation can be written as (see section 5.2.5):

$$Z^{vh} = C\ S_{vh}\ r^2 = \left[\frac{Cr_0^2 N_v}{g_v^r g_h^t}\right]\left[\frac{r^2}{r_o^2}\right]\left[\frac{T_o^{vh} - N_v}{N_v}\right], \text{ where } T_0^{vh} = g_v^r g_h^t S_{vh} - N_v$$

$$= \left[\frac{Cr_0^2 N_h}{g_h^r g_h^t}\right]\left[\frac{r^2}{r_o^2}\right]\left[\frac{g_h^r}{g_v^r}\right]\left[\frac{T_o^{vh} - N_v}{N_h}\right]$$

The third term is simply 1/XDR so that we can write:

$$Z^{vh} = \left[\frac{Cr_0^2 N_h}{g_h^r g_h^t}\right]\left[\frac{r^2}{r_o^2}\right]\left[\frac{xdr^{-1}T_o^{vh} - xdr^{-1}N_v}{N_h}\right]$$

In this case, the first term is the dBZ_o for the HH channel. Thus we can use the dBZ_o for the HH channel to calibrate the cross-channel, if we first adjust the cross–channel noise and power by 1/XDR and then normalize by N_h. The reflectivity calculation assumes that the calibrated XDR value compensates for any differences in the radar constant between the two channels, i.e., we do not need to have separate radar constants for the two channels.

HH+VH Channels

This choice would be used for elliptic transmit polarizations that give comparable return signal in both the co- and cross-channels. The approach is to obtain average autocorrelation functions as follows:

$$T_0 = \frac{T_o^{hh} + xdr^{-1}\ T_o^{vh}}{2}$$

$$R_0 = \frac{R_o^{hh} + xdr^{-1}\ R_o^{vh}}{2}$$

$$R_1 = \frac{R_1^{hh} + xdr^{-1}\ R_1^{vh}}{2}$$

$$R_2 = \frac{R_2^{hh} + xdr^{-1}\ R_2^{vh}}{2}$$

$$N = \frac{N_h + xdr^{-1}\ N_v}{2}$$

These adjusted autocorrelations are then used as per the standard moment processing for calibration with respect to the HH channel.

Case 1V: Fixed Vertical Transmit and Dual Channel Receive- (VV, HV)				
dBZo from VV Channel	**TTY Setup Question Responses**			
Calculate T, Z, V, W from:	**H–Xmt**	**V–Xmt**	**Co–Rcv**	**Cx–Rcv**
VV (co)	——	——	YES	NO
HV (xdr weighting)	——	——	NO	YES
VV+HV (xdr weighting)	——	——	YES	YES

This is the only case for which the calibration constant dBZ_o for the VV channel should be downloaded to the signal processor.

VV Channel (co–pol)

This is the recommended channel for the case of linear polarization. The reason is that for linear polarization, the co-polar channel will have the strongest signal. Processing is identical to a conventional radar.

HV Channel (cross–pol)

This choice would be used for circular or elliptic transmit polarization when most of the return is in the cross-pol channel. Since the algorithm assumes that dBZo is from the co-polar channel, xdr is used to adjust the autocorrelations as follows:

$$T_0 = xdr\ T_o^{hv}$$

$$R_0 = xdr\ R_o^{hv}$$

$$R_1 = xdr\ R_1^{hv}$$

$$R_2 = xdr\ R_2^{hv}$$

$$N = xdr\ N_h$$

These adjusted autocorrelations are then used as per the standard moment processing with dBZo calibrated with respect to the VV channel.

VV+HV Channels

This choice would be used for elliptic transmit polarizations that give comparable return signal in both the co- and cross-channels. The approach is to obtain average autocorrelation functions as follows:

$$T_0 = \frac{T_o^{vv} + xdr\ T_o^{hv}}{2}$$

$$R_0 = \frac{R_o^{vv} + xdr\ R_o^{hv}}{2}$$

$$R_1 = \frac{R_1^{vv} + xdr\ R_1^{hv}}{2}$$

$$R_2 = \frac{R_2^{vv} + xdr\ R_2^{hv}}{2}$$

$$N = \frac{N_v + xdr\ N_h}{2}$$

These adjusted autocorrelations are then used as input to the standard moment processing algorithms with dBZ_o calibrated with respect to the VV channel.

Case 2: Simultaneous Transmit and Receive- STAR (HH, VV) **Case 3: Alternating Transmit Single-Channel Receive (HH, VV)**				
dBZo from HH Channel	**TTY Setup Question Responses**			
Calculate T, Z, V, W from:	**H–Xmt**	**V–Xmt**	**Co–Rcv**	**Cx–Rcv**
HH	YES	NO	—	—
VV (gdr^{-1} weighting)	NO	YES	—	—
HH+VV (gdr^{-1} weighting)	YES	YES	—	—

A fundamental difference between these two cases is that for all standard moment processing choices, the STAR case has double the number of samples as compared to the single-channel alternating case. However, the processing is otherwise identical.

HH Channel

Since the HH channel is directly calibrated this is the recommended choice. Processing is identical to a conventional radar.

VV Channel

In this case, GDR is used to adjust the autocorrelations as follows:

$$T_0 = gdr^{-1}\ T_o^{vv}$$

$$R_0 = gdr^{-1}\ R_o^{vv}$$

$$R_1 = gdr^{-1}\ R_1^{vv}$$

$$R_2 = gdr^{-1}\ R_2^{vv}$$

$$N = gdr^{-1}\ N_v$$

These adjusted autocorrelations are then used as input to the standard moment processing algorithms with dBZo calibrated with respect to the HH channel.

HH+VV Channels

This approach gives the benefit of doubling the number of samples used for the reflectivity calculation.

$$T_0 = \frac{T_o^{hh} + gdr^{-1}\ T_o^{vv}}{2}$$

$$R_0 = \frac{R_o^{hh} + gdr^{-1}\ R_o^{vv}}{2}$$

$$R_1 = \frac{R_1^{hh} + gdr^{-1}\ R_1^{vv}}{2}$$

$$R_2 = \frac{R_2^{hh} + gdr^{-1}\ R_2^{vv}}{2}$$

$$N = \frac{N_h + gdr^{-1}\ N_v}{2}$$

These adjusted autocorrelations are then used as input to the standard moment processing algorithms with dBZo calibrated with respect to the HH channel.

Case 4: Alternating Dual-Channel (HH, VH, VV, HV)				
dBZo from HH Channel	**TTY Setup Question Responses**			
Calculate T, Z, V, W from:	H–Xmt	V–Xmt	Co–Rcv	Cx–Rcv
HH	YES	NO	YES	NO
VH (xdr^{-1} weighting)	YES	NO	NO	YES
VV (gdr^{-1} weighting)	NO	YES	YES	NO
HV (xdr/gdr weighting)	NO	YES	NO	YES
HH+VV (gdr^{-1} weighting)	YES	YES	YES	NO
HV+VH (xdr & gdr weighting)	YES	YES	NO	YES

HH Channel

Since the HH channel is directly calibrated this is the recommended choice. Processing is identical to a conventional radar.

VH Channel

Processing is identical to Case 1H: Horizontal Transmit HV Processing.

VV Channel

Processing is identical to Cases 2&3:STAR and Single Channel Alternating VV Processing.

HV Channel

The weighting in this case uses both xdr and GDR.

$$T_0 = \frac{xdr}{gdr} T_o^{hv}$$

$$R_0 = \frac{xdr}{gdr} R_o^{hv}$$

$$R_1 = \frac{xdr}{gdr} R_1^{hv}$$

$$R_2 = \frac{xdr}{gdr} R_2^{hv}$$

$$N = \frac{xdr}{gdr} N_h$$

These adjusted autocorrelations are then used as input to the standard moment processing algorithms with dBZo calibrated with respect to the HH channel.

HH + VV Channels

Processing is identical to Cases 2&3: STAR and Single Channel Alternating HH+VV Processing.

HV + VH Processing

The weighting here has to correct for both transmitter and receiver effects in order to use the HH channel dBZ_o.

$$T_0 = \frac{\frac{xdr}{gdr} T_o^{hv} + xdr^{-1}\ T_o^{vh}}{2}$$

$$R_0 = \frac{\frac{xdr}{gdr} R_o^{hv} + xdr^{-1}\ R_o^{vh}}{2}$$

$$R_1 = \frac{\frac{xdr}{gdr} R_1^{hv} + xdr^{-1}\ R_1^{vh}}{2}$$

$$R_2 = \frac{\frac{xdr}{gdr} R_2^{hv} + xdr^{-1}\ R_2^{vh}}{2}$$

$$N = \frac{\frac{xdr}{gdr} N_h + xdr^{-1}\ N_v}{2}$$

These adjusted autocorrelations are then used as input to the standard moment processing algorithms with dBZo calibrated with respect to the HH channel.

An example of how this weighted averaging works is given here. Suppose that we want to compute the average of the reflectivities for the VH and HV channels,

$$Z^{hv+vh} = Cr^2\ \frac{S_{hv} + S_{vh}}{2}$$

$$= Cr^2\ \frac{\frac{T_0^{hv} - N_h}{g_h^r g_v^t} + \frac{T_o^{vh} - N_v}{g_v^r g_h^t}}{2} = \frac{Cr^2}{g_h^r g_h^t}\ \frac{(T_0^{hv} - N_h)\frac{g_h^t}{g_v^t} + (T_o^{vh} - N_v)\frac{g_h^r}{g_v^r}}{2}$$

but since $xdr = \frac{g_v^r}{g_v^h}$ and $gdr = \frac{g_v^r g_v^t}{g_h^r g_h^t}$

$$Z^{vh+hv} = \frac{Cr^2}{g_h^r g_h^t} \left[\frac{\frac{xdr}{gdr} T_0^{hv} + xdr^{-1} T_0^{vh}}{2} - \frac{\frac{xdr}{gdr} N_h + xdr^{-1} N_v}{2} \right]$$

$$Z^{vh+hv} = \frac{Cr^2}{g_h^r g_h^t}\ \left[T_0 - N\right] = \left[\frac{Cr_o^2 N_h}{g_h^r g_h^t}\right]\left[\frac{r^2}{r_o^2}\right]\left[\frac{T_0 - N}{N_h}\right]$$

The first term in brackets is precisely dBZ_o for the HH channel. Thus if we average the correlations using the appropriate GDR and xdr weighting as shown above, then the average reflectivity is obtained by using conventional processing with the HH channel dBZ_o.

5.7.12 Thresholding of Polarization Parameters

The thresholding of polarization parameters by the processor eliminates bins with weak or uncertain signals. Note that the thresholding can be disabled if it is desired to see all of the data regardless of the data quality.

All of the polarization parameters are based on power ratios. The RVP7 requires that each power term in a ratio pass a signal-to-noise test similar to the log power test. For example, there are up to four different powers that can be calculated (alternating dual-channel case) so the tests for each of these are:

$$\frac{<|s_{hh}|^2>}{N_h} > N_{thresh}$$

$$\frac{<|s_{hv}|^2>}{N_h} > N_{thresh}$$

$$\frac{<|s_{vv}|^2>}{N_v} > N_{thresh}$$

$$\frac{<|s_{vh}|^2>}{N_h} > N_{thresh}$$

where the linearized threshold that is input as the dB LOG threshold, i.e.,

$$N_{thresh} = 10^{LOGthresh/10}$$

For example, a valid LDRH requires both a valid S_{hh} and a valid S_{vh}. The parameters RHOH and PHIH have the same requirement since they are the magnitude and phase of the cross-correlation function which is based on S_{hh} and S_{vh}.

There are two exceptions:

ZDR

ZDR requires that both S_{hh} and S_{vv} pass the signal-to-noise tests noted above. However, ZDR can be additionally thresholded by any of the other threshold parameters (LOG, SIG, SQI, CSR) similar to a standard moment. See section 5.3 for a description of the standard moment thresholding.

PHIDP for single channel alternating case

PHIDP requires that both S_{hh} and S_{vv} pass the signal-to-noise tests noted above. In the single channel alternating case, PHIDP must also satisfy the additional test that the Doppler velocity at the range bin must be valid, i.e., not thresholded by its own criteria. This is because the algorithm for PHIDP in this case essentially subtracts the phase change due to the Doppler velocity. If the Doppler velocity is uncertain, the algorithm cannot produce reliable results.

5.7.13 Calibration Considerations

Polarization systems require additional calibration as compared to conventional systems. There are three aspects to the calibration:

- dBZ_o measurement in both channels for dBZ and dBT calibration.
- GDR measurement for ZDR calibration.
- xdr measurement for LDR calibration.

These are discussed below.

dBZ_o Calibration for dBZ

The RVP7 supports separate calibration of both polarization channels. Measurement of dBZ_o for each channel of a dual polarization system is identical to the conventional radar case described in Section 5.4. Note that for a single-channel switching system, the only difference between the horizontal and vertical signal paths occurs after the high power switch, i.e., differential insertion loss of the switch itself and any differential insertion loss of the waveguides and feed after the switch. This means that for single-channel switching systems it may be sufficient to calibrate at one polarization and then adjust the calibration of the other channel by the differential gain GDR (see below).

GDR Calibration for ZDR

GDR is the relative between the co-polarized channels including both transmitter and receiver gain, i.e.,

$$GDR = 10\ LOG\ \frac{g_v^r g_v^t}{g_h^r g_h^t} \qquad \text{and} \qquad gdr = \frac{g_v^r g_v^t}{g_h^r g_h^t}$$

GDR is input into the processor as a dB value. However, for analyses in this chapter, the linear gdr value is sometimes more convenient.

In principle, if dBZ_o could be calibrated perfectly in both channels, measurement of GDR would not be required. In practice, this is not possible because dBZ_o cannot be calibrated to an absolute accuracy sufficient for ZDR, i.e., to 1/16th of a dB. Therefore, the RVP7 uses the GDR approach.

Since GDR includes both transmitter and receiver differential gains, accurate calibration requires that an actual target be observed. One way to do this is as follows:

- Set the GDR to be 0 dB using your application software (e.g., for SIGMET IRIS systems in the setup utility RVP section). Disable clutter filtering for ZDR in either your application software (by selecting filter 0) or explicitly in the RVP7 TTY setups mp section.
- Place the antenna at 90 degrees elevation (vertical incidence) during moderate to heavy rain. The melting layer should be at a height that is well above the recovery zone of the T/R and in the antenna "far zone". A melting layer higher than 2 km is suggested, but the specific characteristics of the radar should be considered.

- Collect ZDR data at vertical incidence while the antenna is rotating in azimuth.
- Use a separate application program to average the ZDR values around a full 360 degrees at each range bin (height). Generate a plot of 360-average ZDR vs height.
- You should observe that the average ZDR values in regions of strong signal (>20 dB SNR) below the bright band are approximately constant with height. This is the value that should be used in your application software for GDR.
- Enter the value and repeat the calibration to verify that the average ZDR is now 0 dB.

The rationale for this approach is as follows. When viewed at vertical incidence, rain should have a ZDR of 0 dB since the drops will all appear circular. The reason for averaging over 360 degrees is to cancel-out effects from sidelobe contamination from nearby ground targets and other artifacts of the antenna/feed/radome system. For example the radome may have an obstruction light on the top. Some of these artifacts can be minimized by assuring the the weather targets are strong, i.e., heavy rain is preferred for this calibration.

XDR Calibration for LDR

XDR is the relative gain in dB between the co- and cross-receiver channels for LDR measurements. Analogous to GDR, it is defined as the dB value of the ratio of the vertical to horizontal receiver gains, i.e.,

$$XDR = 10\ LOG\ \frac{g_v^r}{g_h^r} \qquad \text{and} \qquad xdr = \frac{g_v^r}{g_h^r}$$

Three techniques for calibration of XDR are discussed. It is recommended for the transmitter to be off for all of these methods.

- 1. **Solar method**
  Use the sun to measure LDR. The measured value of LDR is then the XDR offset. LDR should be measured in fixed mode for both LDRH and LDRV. The values should be reciprocal (e.g., +1 dB and –1 dB). Use the average of the absolute value if they are not precisely reciprocal (e.g., for +1.4 and –1.2 use 1.3). Finally after inputting the XDR value, retest to verify that the sun has been properly corrected to have zero LDR.
- 2. **Signal generator method with connection to waveguide**
  Connect a signal generator with a splitter to both channels and measure XDR directly. This does not account for any effects that are before the coupler (e.g., waveguide, feed, radome, antenna gain).
- 3. **Linear feed horn remote radiator method**
  Use a calibrated linear feed horn with an RF source located several hundred meters from the radar. Maximize the H channel return and measure the response using the RVP7 pr command "Filtered" power in the "Primary Channel". Now rotate the feed horn to vertical and maximize the power in the "Secondary Channel". The difference in dB is XDR. Note that signal multi–path effects could bias the results from this technique.

In all cases it is recommended that for the calibration, XDR be set to 0 dB in the application user software and that the RVP7 TTY setups be configured as follows:

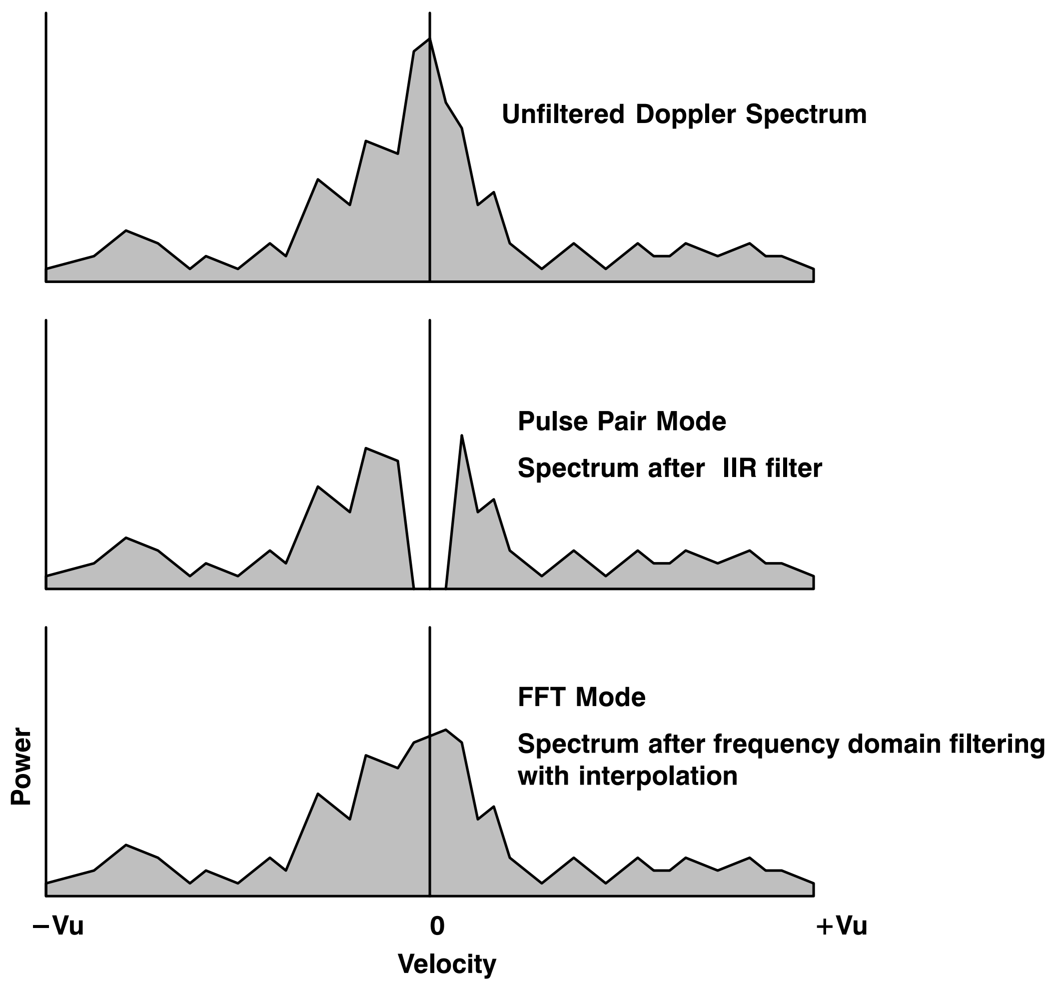
- Noise correction enabled for LDR and noise sample taken prior to the measurements (with care not to sample with a test signal turned–on or while looking at the sun).
- Clutter correction disabled for LDR.

5.8 FFT Mode

5.8.1 Overview

The RVP7 can perform FFT processing of the I and Q time series. This is indicated by the inset box in Figure 5–1. The major difference between FFT and pulse pair processing is the way in which clutter filtering is performed. The pulse pair mode uses a time domain IIR filter while the FFT mode uses a frequency domain filter. The advantage of the FFT approach is that it is less destructive to overlapped weather (near zero velocity) than the IIR filter since the clutter filter algorithm attempts to interpolate over the weather (see Figure 5–8). This results in more accurate estimates of velocity, width and clutter correction. Because the clutter correction is more accurate, the resulting reflectivity estimates are more accurate.

Figure 5–8: Comparison of Pulse Pair and FFT Clutter Filters



5.8.2 FFT Implementation

Figure 5–9 shows a data flow diagram for the FFT processing. In the example, 50 pulses of I and Q are processed. Recall that the I and Q samples refer to a single range bin with an I and Q sample generated for each pulse.

Sampling Division

The first step of processing is to split the I and Q values into two groups. The FFT algorithm requires the number of samples to be a power of two. This is not very convenient for radar sampling since it effectively quantizes the scan rates to be powers of two. The processing by the RVP7 allows an arbitrary number of pulses to be handled by splitting the samples into two groups — the first 2^N and last 2^N pulses. An FFT is performed on each group and then the results are averaged. This allows an arbitrary number of samples to be processed. The maximum number of samples is 255 which corresponds to performing FFT's on two groups of 128 pulses with 1 sample of overlap.

If the requested number of pulses is exactly a power of 2, then the samples are not split and only a single FFT is done.

Window and FFT

After splitting the samples, a weighting function, or window, is applied. The FFT algorithm is then applied and the magnitude squared of each component is calculated to yield the Doppler power spectrum. Figure 5–10 shows the response of the Doppler power spectrum to a ground clutter target for each of the three available FFT windows. The schematic Doppler spectra in the example do not show the details of the side lobe structure of the impulse response (see Oppenheim and Schafer, 1975 p. 243–245).

- **Rectangular Window** — This is really the no-window case since it is equivalent to multiplying each sample by 1. This window has a sinx/x type impulse response, i.e., the impulse response is very narrow which is good, but the side lobes of the Doppler spectrum are rather high — ~20 dB near the peak. Thus it is not the best choice for high performance clutter cancelation since the high sidelobes will mask weak weather targets.

- **Hamming Window** — This common window is well matched to magnetron systems. The peak of the impulse response is broader than that for the rectangular window, but it has –40 dB peak-to-noise sidelobes. Weak weather targets above this level will not be obscured by these sidelobes. The phase noise of a magnetron system (~1 degree) will typically be slightly larger than the sidelobes caused by the window which means that the window is well matched to the magnetron performance.

Figure 5–9: FFT Processing — 50 pulse example

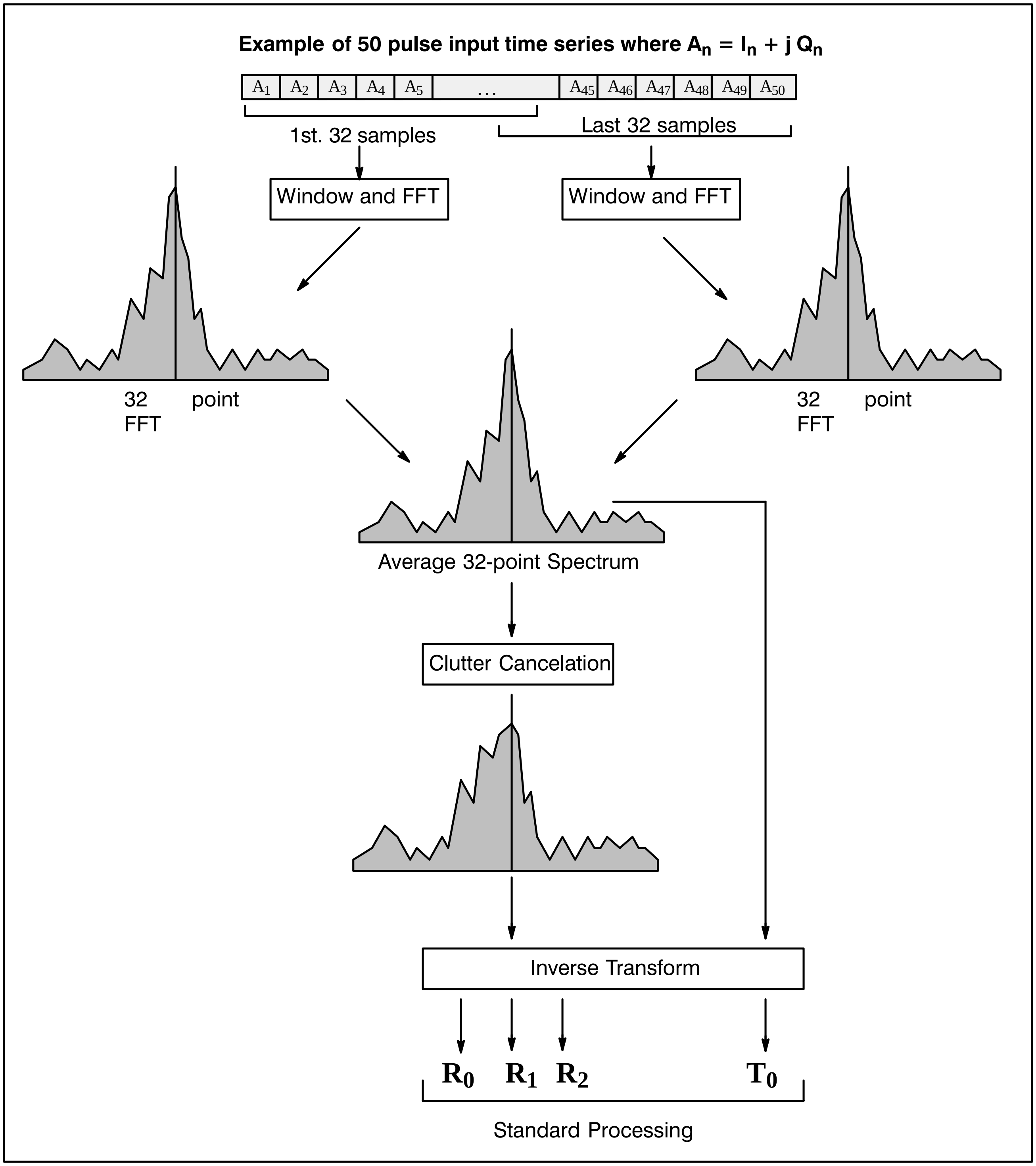
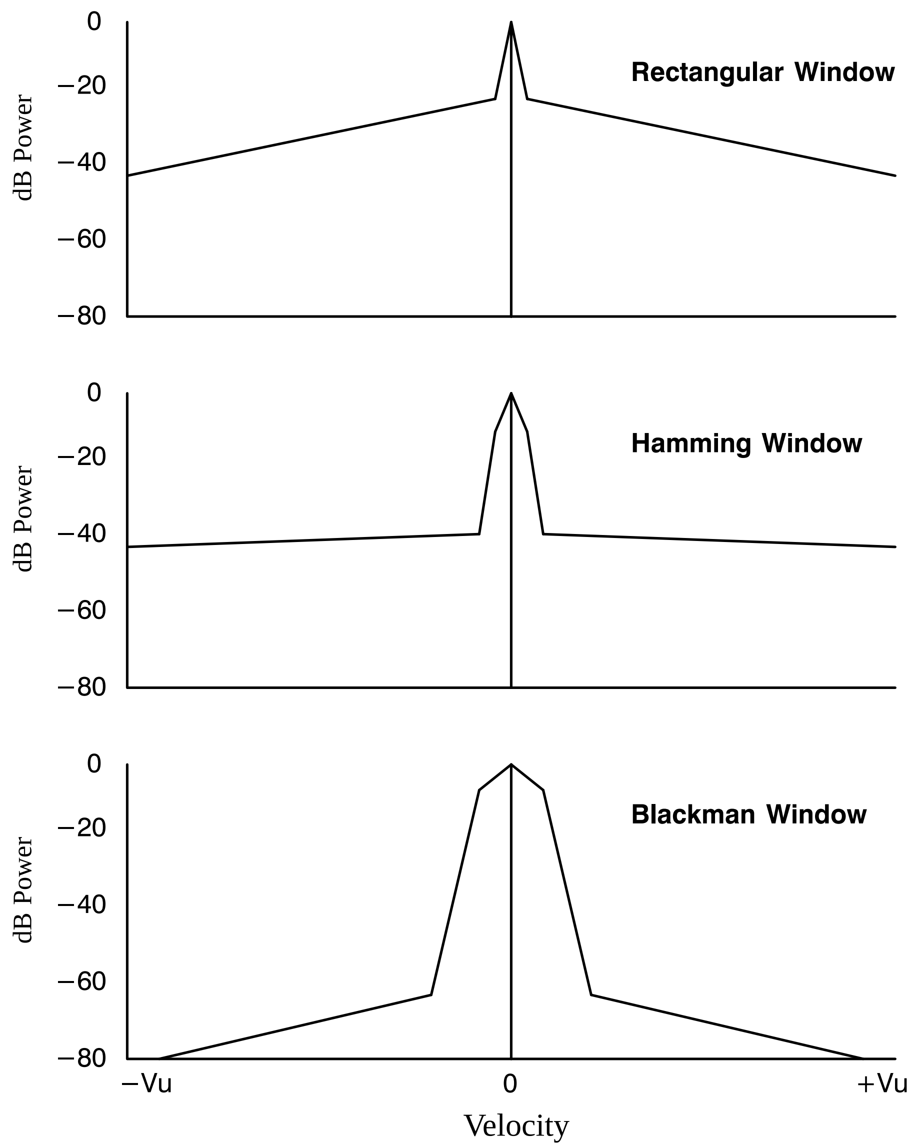


Figure 5–10: Effect of Windowing on FFT Response to Ground Clutter



- **Blackman Window** — This is the most aggressive window for clutter cancelation and is appropriate only for Klystron systems that can achieve very low phase noise (~0.1 degree). The peak is the broadest of all the windows, but the side lobes are the lowest. It is not recommended for magnetron systems except for performance testing.

FFT Averaging

The power spectrum from the first group of samples is averaged with the power spectrum from the last group of samples. Note that if the total number of samples is exactly a power of two, then this step is skipped. Averaging the two power spectra from the overlapping sample groups effectively captures the information from all of the samples. The result is a smoother power spectrum than weather of the individual spectra.

Ground Clutter Cancellation

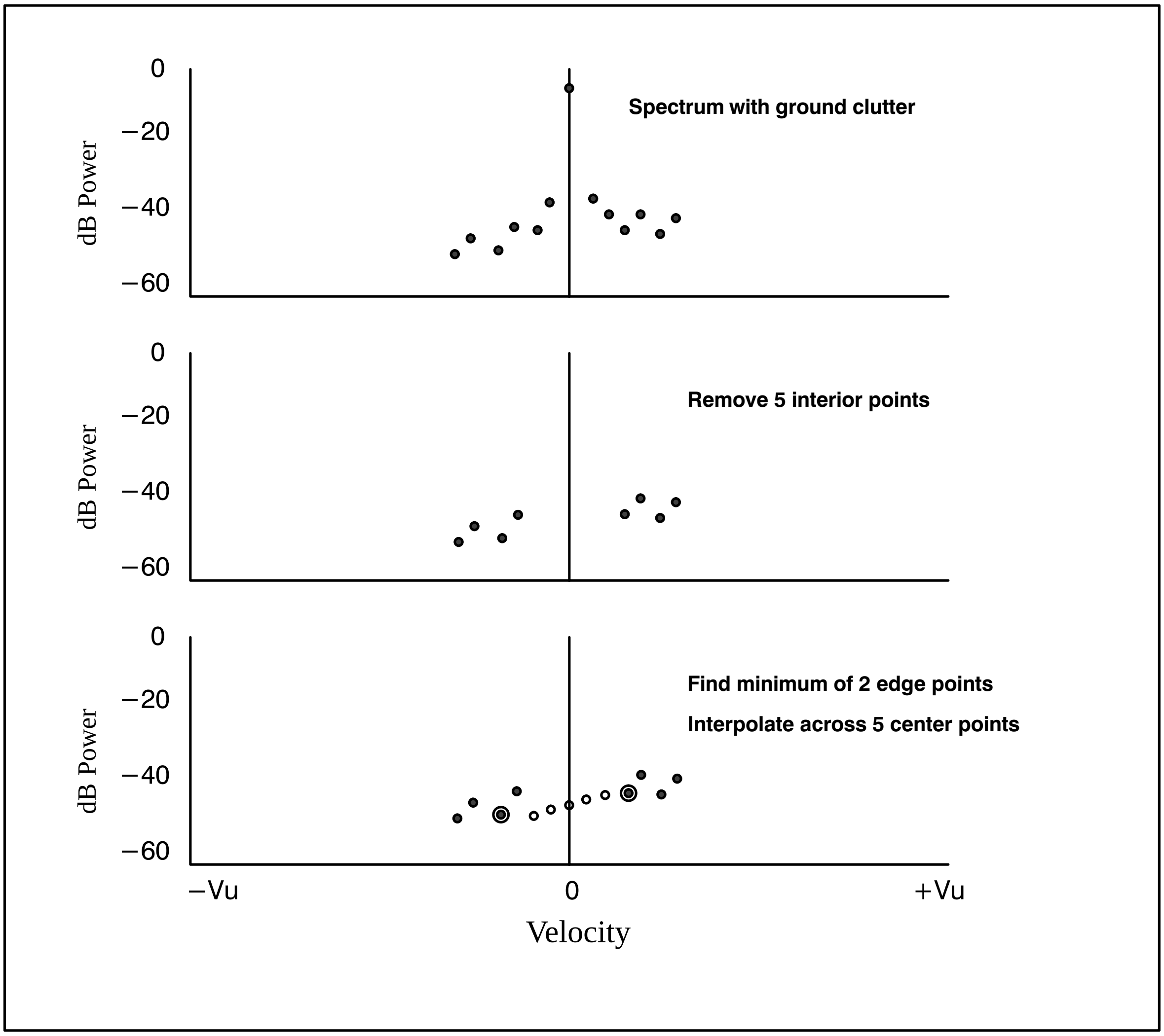
The ground clutter cancelation is performed directly in the frequency domain by either of two techniques:

- Fixed width filters- Interpolation over a fixed but selectable width.
- Variable width filters- Interpolation over an adaptive width that adjusts for the width of the clutter.

In general, SIGMET recommends the use of the adaptive filter approach since there is adaptive compensation for different antenna rotation rates, and minimal if any damage or bias is introduced in the case of overlapped weather echoes.

See Section **3.3.3** on how to configure these two types of filters and provides algorithmic details.

Figure 5–11: Example of Fixed–Width Frequency Domain Clutter Filter



Fixed Width Filters

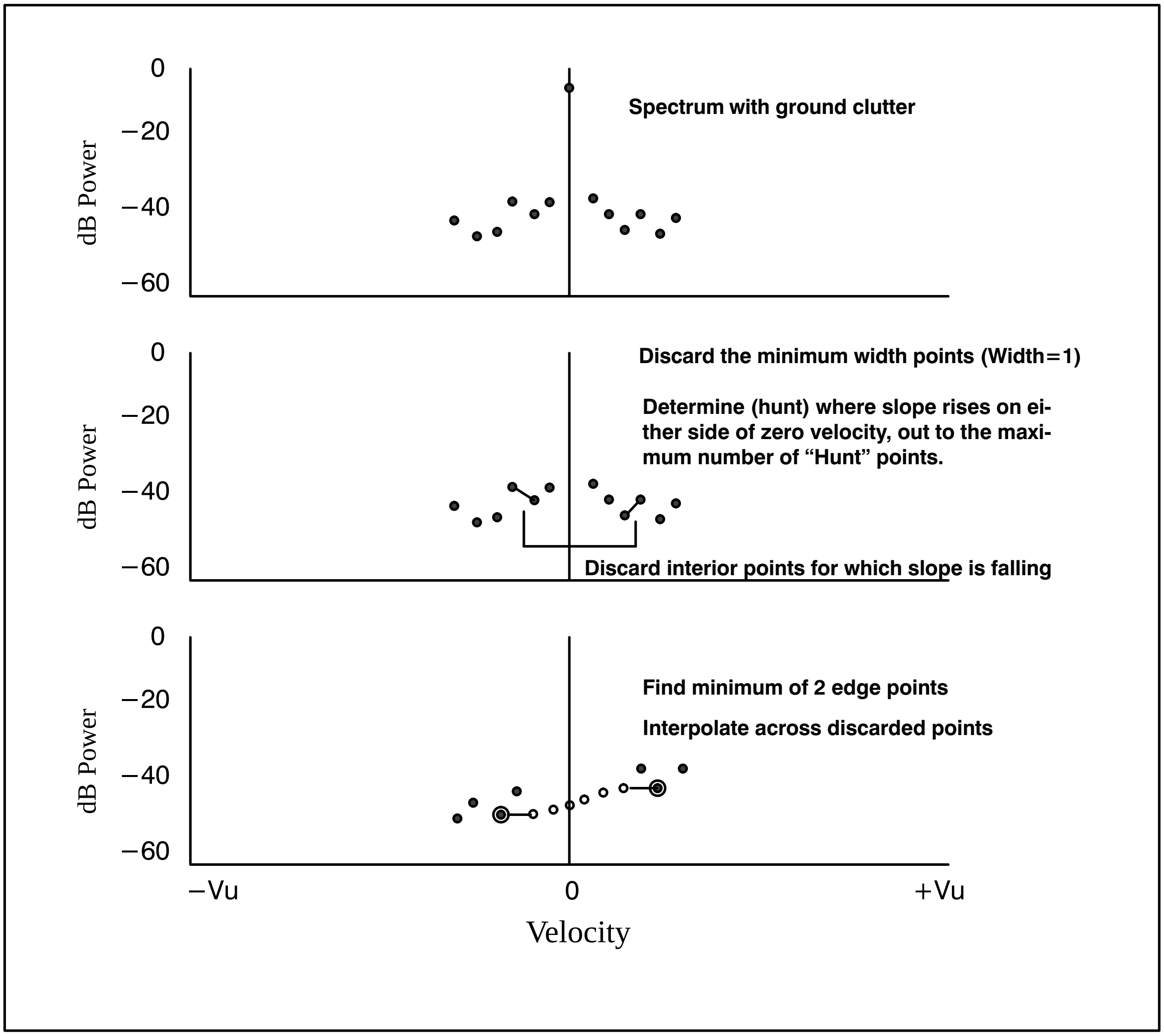
An example is shown in Figure 5–11. In general, the technique discards a selectable number of center points. The algorithm then takes the minimum value of a selectable number of edge points next to the discarded points. The minimum value on each side is used to interpolate across the points that were removed. In the example, the number of center points removed is 5 and the number of edge points used is 2.

This procedure preserves the noise level and/or overlapped weather targets. The result is that more accurate estimates of dBZ are obtained. In extreme cases when the weather spectrum is very narrow, there can still be some attenuation of weather if a broad filter is selected.

For reference, the corresponding setup for this filter in the **mf** setup for "Spectral Clutter Filters" would be (see Section **3.3.3**) :

```
Filter #N – Type:0 (Fixed)     Width: 3     Edgepoints: 2
```

Figure 5–12: Example of Variable Width Frequency Domain Clutter Filter



Variable Width Filters

These function similarly to the fixed with filters except that now the width of the discarded points is allowed to change in response to clutter of different widths. The edge of the clutter is signaled by a rising slope (moving outward from zero velocity). A minimum number of points (”Width”) is always discarded, and there is a maximum number of additional points (”Hunt”) over which the algorithm will hunt for the rising slope. In the event that the rising slope is not found after searching the maximum number of points, then the width is fixed to the maximum.

For reference, the corresponding setup for this filter in the **mf** setup for "Spectral Clutter Filters" would be (see Section **3.3.3**) :

```
Filter #N – Type:1 (Variable)   Width: 1   Edgepoints: 2   Hunt:  3
```

The optimal filter in most cases can be determined by experimenting to find the minimum width fixed-width filter that eliminates all clutter (e.g., based on a PPI display of velocity or dBZ on a clear day). Then using a variable width filter with (minimum) "Width" 1 or 2 and a "Hunt" that extends this to the width of the fixed width clutter filter. For example if the fixed width clutter filter that just removes all of the clutter is configured as follows:

```
Filter #N – Type:0 (Fixed)     Width: 4     Edgepoints: 2
```

"Width: 4" means that the center point at zero velocity, plus three additional points on either side of zero velocity are removed. Therefore a suggested optimal variable width filter would be:

```
Filter #N – Type:1 (Variable)   Width: 1    Edgepoints: 2   Hunt: 3
```

This filter can be checked and tuned to be more aggressive by either increasing the Width to 2 and/or the Hunt to 3.

Inverse Transform

After clutter removal, an inverse DFT (Discrete Fourier Transform) is performed to obtain the autocorrelations R0, R1 and R2 (optional). The total power T0 in the unfiltered power spectrum is also computed by summing the spectrum components. Thus the final output of the FFT approach is identical to the pulse pair approach except that the clutter filtering is performed in the frequency domain.

5.9 Random Phase 2nd Trip Processing

5.9.1 Overview

Second trip echoes can be a serious problem for applications when the radar is operated at high PRF (e.g., >500 Hz). Second trip echoes are caused by the range aliasing of targets. They appear as false echoes on the display, usually elongated in the radial direction. On Klystron systems they will have valid Doppler velocities. On magnetron systems, the Doppler velocities are not valid, but the noise from the 2nd trip echoes can obscure valid first trip velocity information.

The RVP7 has optional random phase processing for the filtering and recovery of second trip echoes. Details of the technique are proprietary to SIGMET, Inc. However, the general principle is described here, along with a discussion of the various configuration options to optimize the algorithm performance.

The information that is used to separate the first and second trip echoes is the phase. For a magnetron radar, the phase of each pulse is different. This means that when 1st. and 2nd trip echoes are received simultaneously, the phase of the first trip return is different from the phase of the second trip return. For a magnetron radar, the RVP7 measures the phase of the transmitted pulse and the phase locking is done digitally as opposed to the traditionally locking COHO. For a Klystron radar, the phase is controlled by the RVP7 via a digital phase shifter that is precisely calibrated. Typically the Klystron COHO is phase shifted so that each transmit pulse has a different phase. The sequencing is controlled by the RVP7.

5.9.2 Algorithm

Figure 5–13 shows a schematic of the data processing for random phase. The figure shows the Doppler spectra for the 1st. and 2nd trip in the various processing stages. The vertical scale is in dB and the horizontal scale is velocity. In this example, the second trip echo is shown as being stronger than the first trip echo (usually the reverse is true).

Ideal 1st and 2nd Trip Echoes

The ideal 1st and 2nd trip echoes represent the echoes as they would appear individually. The ideal 1st trip echo is the echo that would be measured if there were no 2nd trip echo interference. The ideal 2nd trip echo represents what would be measured if there were no 1st trip echo interference. If there is no interference from the other trip, a standard Klystron system can measure the ideal spectra, but there is no way to know whether the echoes are in the 1st or 2nd trip.

Raw 1st and 2nd Trip Echoes

This figure shows how the echoes from the first trip and second trip interfere with each other. For the case of a standard magnetron system, the first trip echo is coherent, while the second trip echo is incoherent (white noise) since the phase of the second trip echo is random. This is because the receiver is phase locked only to the first trip.

Another way to implement a magnetron system is to let the COHO free-run (rather than phase locking to the transmit pulse), measure the phase of each transmit pulse and digitally correcting for the transmit phase. Using this digital phase locking technique, the RVP7 can phase lock or "cohere" to either the first or the second trip.

Using this technique alone, it is possible to distinguish between 1st and 2nd trip echoes for the case when the echoes are not overlapped. In other words, the echoes will appear as the idealized 1st and 2nd trip echoes. This range de-aliasing effectively doubles the range of the radar. The problem is that when echoes are overlapped, the noise contamination from the stronger echo will make it impossible to measure the weaker echo. This is illustrated in the figure. Thus if the first trip echo has a good signal-to-noise ratio of 10 dB, then the 2nd trip echo will have a signal-to noise-ratio no better than –10 dB. This is the fundamental problem with using phase alone to separate the 1st and 2nd trip echoes.

Filtered 1st and 2nd Trip Echoes

Since the strong echo generates noise that obscures the weaker echo, the approach used in the RVP7 is to filter the echo from the other trip — the whitening filter. This is shown in the figure. The adaptive whitening filter removes both the clutter and the weather. All of the phase information for the other trip is then contained in the white noise portion of the spectrum. Note that the phase information under the coherent echo that is removed will be dominated by the coherent echo, i.e., the other trip phase information will be contaminated. For this reason, the filtering should effect as small a region of the spectrum as possible.

5.9.3 Tuning for Optimal Performance

The Random Phase algorithms are controlled by the same collection of setup and operational parameters that apply to all of the other processing modes, e.g., choice of sample size, clutter filter, angle sync, calibration, etc. However, a few parameters are special to Random Phase mode, and these are described below.

Secondary SQI Threshold

In standard Doppler processing, an SQI threshold is normally not applied to Reflectivity data because it would cause those data to be rejected in regions of high spectral width. In Random Phase mode we need to relax this convention because reflected power can only be assigned to a particular trip when it is coherent within that trip. Incoherent echoes, regardless of their strength, can not be placed into either trip.

Thus, an SQI threshold is required to qualify Reflectivity data in Random Phase mode. The RVP7 defines a secondary SQI threshold SQI_2 which is computed from the standard threshold value simply as:

$$SQI_2 = Offset + (Slope \times SQI)$$

Where $Slope$ and $Offset$ are the Random Phase SQI threshold parameters defined in the **Mf** setup section. The factory default values are $(Slope = 0.50)$ and $(Offset = -0.05)$, i.e., the secondary threshold is a little less than half of the standard value. The Random Phase algorithms check whether the SQI of each recovered trip is less than the secondary SQI threshold, and if so, the LOG portion of the data are rejected. This SQI test is necessary for a clean LOG picture, but we need to use a more permissive (lower) threshold value than would usually be applied to the Doppler data alone.

The *Slope* and *Offset* values should be adjusted so that the density of speckles in Random Phase LOG data is approximately the same as the density of speckles in FFT velocity data for a given primary SQI value. You may then adjust the primary SQI threshold to achieve the appropriate tradeoff of speckles vs. sensitivity for your system in all modes of operation. Even with proper adjustment, it is normal for Random Phase *dBZ* and *dBT* data to show "holes" in regions of weather that have high turbulence or shear. These dropouts will usually match up with similar gaps in the velocity and width data, both of which are traditionally thresholded by SQI.

Maximum Power Ratio Between Trips

The adaptive filtering that is performed on the data for each trip greatly extends the visibility of a weak echo that is overlapped with a much stronger one. In practice, the filtering process is often able to remove 25-35dB of dominant power in order to reveal a much weaker echo in the other trip. The performance depends on many factors, primarily the spectral width of the dominant echo, and the overall stability of the radar system.

The difficulties of removing a dominant "other trip" echo from a weather signal are analogous to the challenge of removing a dominant clutter target from that same signal. In both cases we are trying to extract a weak weather signature using a filtering procedure that relies on the spectral confinement of the stronger signal. The RVP7 already has a parameter that can be adjusted to control sub-clutter visibility, i.e., the Clutter-to-Signal Ratio (CSR). Just as the CSR applies to the clutter filters, it can likewise be used to place similar limits on the depth of visibility of the adaptive filters.

As an example, suppose that the RVP7 is operating in Random Phase mode at a PRF of 1500Hz, and is observing widespread weather having uniform intensity in both the first 100Km trip and the second 100Km trip. If the CSR were set too conservatively at only 15dB, then the algorithm would generally be blind to second-trip weather in the range interval from 100km to 117.8km. The explanation for this can be found in the $1/r^2$ geometric correction for weather echo intensity. At ranges less than 17.8km, the first trip weather would generally dominate the second trip weather by more than 15dB. Thus, the initial 17.8km ring of second trip data would be rejected by the CSR criteria. However, if the CSR were increased to 30dB, then the size of this missing ring would be reduced to only 3.2km.

If the CSR is set too low you will notice an abrupt ring of missing data in the beginning of the second trip. If set too high, there will be speckles and other spurious effects within this same interval. The optimum setting should strike a balance between these two effects.

R1 vs. R2 Algorithms

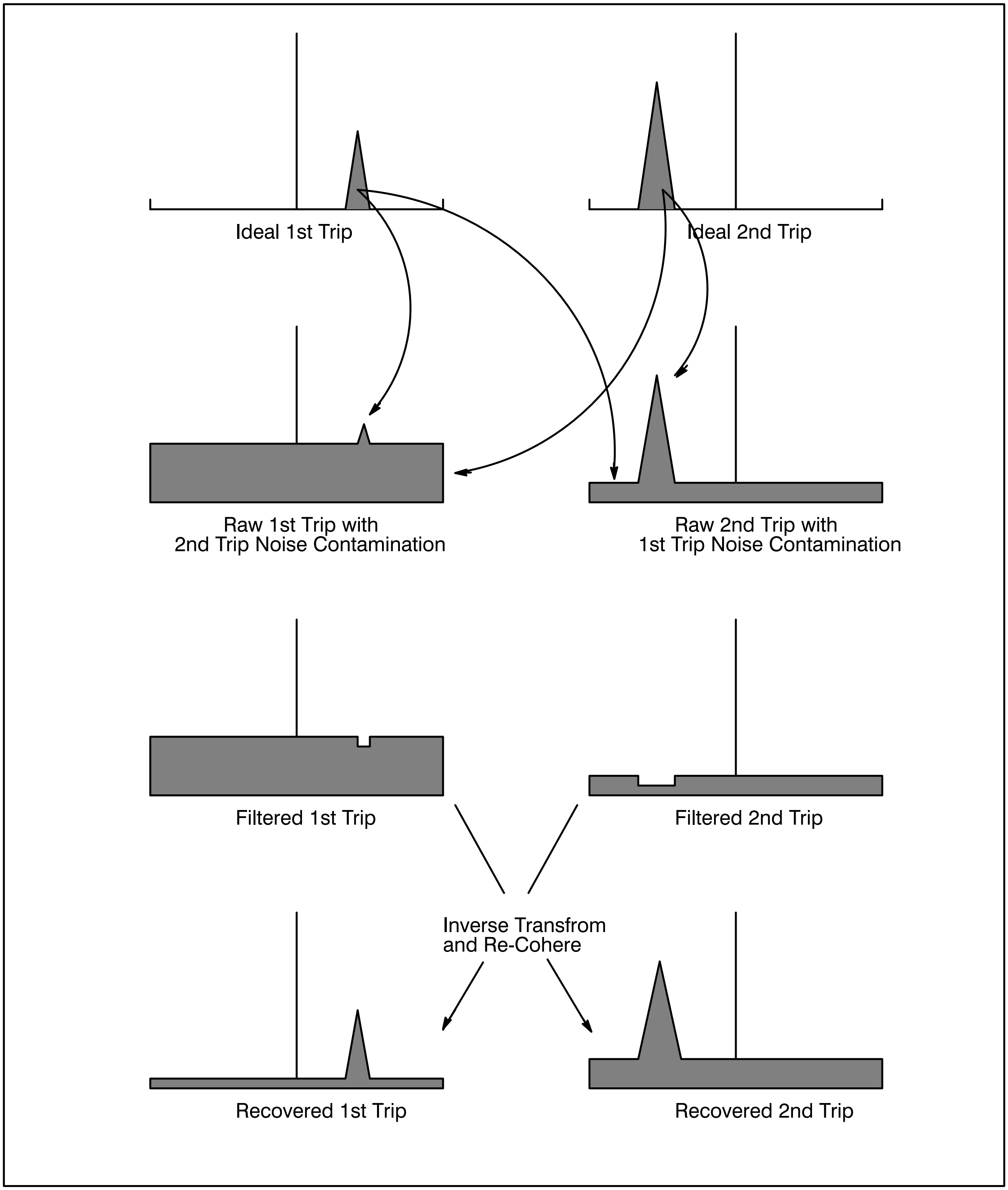
The Random Phase algorithms for adaptive filtering and separation of trips relies on having the best possible information about the weather's SNR and spectral width. Thus, the "R2" Doppler algorithms are always used, regardless of the setting of the R1/R2 flag in the user's operational parameters.

Random Phase and Dual PRF

The random phase processing works seamlessly with the dual PRF processing to provide advanced range and velocity ambiguity resolution. Both the first and 2nd trip echoes can be recovered and displayed to a maximum range of 2X the unambiguous range corresponding to the high PRF.

For optimum performance, the 2D 3x3 speckle filter should be used to smooth the 2nd trip seams that occur for each ray. In fact, this smoothing of the 2nd trip seam makes the dual PRF random phase mode work even better than the single PRF random phase.

Figure 5–13: Random Phase Processing Algorithm



5.10 Signal Generator Testing of the Algorithms

This section describes a variety of IF signal generator tests that can be used to verify correctness of the RVP7 processing algorithms. These tests are routinely performed at SIGMET whenever new algorithms and/or major modes are added to the processor. We have include a few of the test descriptions here so that they can be used by customers who need to debug their systems, or who want to better understand how they work. Additional tests for receiver sensitivity and dynamic range can be found in Appendix E.

5.10.1 Linear Ramp of Velocity with Range

Suppose that a continuous-wave IF waveform has an instantaneous frequency $f(t)$ in Hertz (cycles/sec). Consider a range bin located at time τ_{bin} within a set of pulses that are separated by $\tau_s = 1/PRF$. The phase measured at that bin on the n^{th} pulse will be the integral of the frequency within that pulse starting from range zero (since the RVP7 is phase locked to range zero):

$$\Phi_n = \int_{n\tau_s}^{n\tau_s + \tau_{bin}} f(t)dt$$

If we assume that the input frequency is a linear Frequency Modulation (FM) at the rate of M cycles/sec/sec on top of a base frequency T_o, then:

$$\Phi_{n+1} - \Phi_n = \int_{(n+1)\tau_s}^{(n+1)\tau_s + \tau_{bin}} (T_o + Mt)dt \quad - \int_{n\tau_s}^{n\tau_s + \tau_{bin}} (T_o + Mt)dt \quad = \quad (M\tau_s)\tau_{bin}$$

which, remarkably, is independent of both T_o and n. Thus, a linear FM input signal produces a fixed (I,Q) phase difference from pulse-to-pulse at any given range. The magnitude of the phase difference is proportional to the range, and the slope is $(M\tau_s)$ cycles for each second of delay in range. For example, if the test signal generator is sweeping 100KHz every two seconds, then the velocity observed at a range of 300km at 250Hz PRF will be:

$$\Phi_{n+1} - \Phi_n = \left(\frac{100\ KHz}{2\ \mathrm{sec}}\right) \times \left(\frac{1}{250}\mathrm{sec}\right) \times (300\ km) \times \left(\frac{6.6\ \mu\ \mathrm{sec}}{1\ km}\right) = 0.40\ cycles$$

We would thus observe a velocity of $(0.8 \times V_u)$ at 300km, where V_u is the unambiguous Doppler velocity in meters/sec. Note that these phase difference calculations have made no assumptions about the RVP7 processing mode, and thus are valid in all major modes (PPP, FFT, DPRT, RPH), as well as in all Dual-PRF unfolding modes.

Interestingly, this simple FM signal generator will also produce valid second trip velocities that can be seen during Random Phase processing. This follows from the above analysis because we've never assumed that τ_{bin} was smaller than τ_s, i.e., it is fine for the range bin to be located in any higher-order trip.

5.10.2 Verifying PHIDP and KDP

The PHIDP and KDP processing algorithms can be tested using CW signal sources at IF. In the alternating-transmitter single-receiver case, a single FM signal generator is modulated with an RVP7 polarization select line so that slightly different frequencies are generated for the H and V pulses. A maximum FM depth of several kilohertz is all that is required. In the dual-receiver case, two (unmodulated) signal generators are used for each of the H and V intermediate frequencies, and one or the other is detuned slightly from its correct center frequency. In either case the frequency difference that produces a KDP value of 1.0 degree/km will be:

$$\left(1.0 \ \frac{degree}{km} \right) \times \left(\frac{1}{360} \ \frac{cycles}{degree} \right) \times \left(299792 \ \frac{km}{second} \right) \ = \ 833 \frac{cycles}{second}$$

5.10.3 Verifying RHOH, RHOV, and RHOHV

These three terms measure the normalized cross-channel covariance in a polarization radar. They all are computed in essentially the same way having the form:

$$RHOAB \ = \ \frac{< s_A^n \ s_B^n * >}{\sqrt{< s_A^2 > < s_B^2 >}}$$

Where the s_A^n and s_B^n are complex (I,Q) vectors from two receiver channels A and B, and "<>" denotes expected value. This suggests that some form of amplitude modulation (AM) of the input signal might be helpful.

Suppose that the s_A^n and s_B^n samples are coming from two signal generators installed on a dual-receiver system, and that only the B-Channel is AM modulated so that:

$$| s_A^n | \ = \ \{ S_A, S_A, S_A, S_A, S_a \ ... \} \quad , \quad | s_B^n | \ = \ \{ S_B, 0 \ , S_B, 0 \ , S_B \ ... \}$$

Then the above estimator reduces to:

$$RHOAB \ = \ \frac{(\frac{1}{2}) \, S_A \, S_B}{\sqrt{S_A^2 \times (\frac{1}{2}) \, S_B^2}} \ = \ = \ 0.707$$

A simple way to create these data is to set the A-Channel siggen for 95% AM depth, and use a sinusoidal modulation source of, perhaps, 400Hz. The reason for not choosing 100% depth is that we would loose the Burst phase reference when the amplitude became smallest. The 26dB reduction in S_B is a close enough approximation to zero in the above formula.

If we now observe the two receive channels with the RVP7 at a PRF of 800Hz, we will see the various RHOAB terms varying with range; reaching a high value of 1.00, and a low value of 0.707. The plots will be nearly stationary on the **ascope** screen because the PRF is almost precisely twice the modulation rate (though they are free-running relative to each other).

Adjusting the amplitude of either signal generator will not affect the ρ terms, but it will have an interesting effect on SQI. If (T,Z,V,W) are being computed from both channels combined, then the SQI is:

$$SQI \ = \ \frac{S_A^2}{S_A^2 + (\frac{1}{2}) \, S_B^2}$$

If we solve this equation for SQI=0.5 we find that the individual S_A terms must have twice the power of the individual S_B terms. This can be checked by adjusting either signal generator until the minimum plotted SQI is 0.5, and then verifying that the average *H* and *V* powers are identical; or, equivalently, that *ZDR*, *LDRH* and *LDRV* are zero.

The linear FM ramp described in Section 5.10.1 can also be used as a test of RHOAB in a dual-receiver system. With one siggen modulated and the other fixed, one receive channel will appear to be rotating relative to the other. If the FM modulation is such that *1/N* of a full revolution occurs per pulse at a given range, then if the sample size is *N* pulses we will observe $RHOAB = 0$ at that range. In fact, the plot of $RHOAB$ will show a characteristic $sin(x)/x$ behavior as a function of range.