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# Weighting in Digital Synthetic Aperture Radar Processing

Alan Di Cenzo

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16. Abstract

Weighting is commonly employed in SAR processing to reduce the sidelobe response at the expense of peak center response height and mainlobe resolution. The weighting effectiveness in digital processing depends no only on the choice of weighting function, but on the fineness of sampling and quantization, on the time bandwidth product, on the quadratic phase error, and on the azimuth antenna pattern. This paper reports the results of simulations conducted to uncover the effect of these parameters on azimuth weighting effectiveness. In particular, it is shown that multilook capabilities of future SAR systems may obviate the need for consideration of the antenna pattern, and that azimuth time-bandwidth products of over 200 are probably required before the digital results begin to approach the ideal results.

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# ABSTRACT

Weighting is commonly employed in SAR processing to reduce the sidelobe response at the expense of peak center response height and mainlobe resolution. The weighting effectiveness in digital processing depends not only on the choice of weighting function, but on the fineness of sampling and quantization, on the time bandwidth product, on the quadratic phase error, and on the azimuth antenna pattern. This paper reports the results of simulations conducted to uncover the effect of these parameters on azimuth weighting effectiveness. In particular, it is shown that multilook capabilities of future SAR systems may obviate the need for consideration of the antenna pattern, and that azimuth time-bandwidth products of over 200 are probably required before the digital results begin to approach the ideal results.

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## I. INTRODUCTION

Weighting is often applied to the frequency transfer function in SAR processing to reduce both the integrated sidelobe ratio (ISLR) and the peak sidelobe ratio (PSL) of the impulse response. The procedure may be summarized as follows [1], [2]. Let a linear FM signal be received as

$$s(t) = \exp\left\{j\,\frac{b}{2}\,t^2\right\} \qquad \text{for } -T/2 \le t \le T/2 \tag{1}$$

Then the impulse response, or matched filter output, is given at time $\tau$ by

$$\begin{aligned} \emptyset_{ss}(\tau) &= \int_{-\infty}^{\infty} s(t)s^*(t+\tau)dt \\ &= \int_{-T/2}^{T/2-\tau} \exp\left\{j\,\frac{b}{2}\,t^2\right\}\exp\left\{-j\,\frac{b}{2}\,(t+\tau)^2\right\} dt \end{aligned} \tag{2}$$

Thus

$$\emptyset_{ss}(\tau) = \begin{cases} \dfrac{T\sin\left(\frac{b}{2}T\tau - \frac{b}{2}\tau^2\right)}{\frac{b}{2}T\tau} & |\tau| < T \\ 0 \text{ elsewhere} \end{cases} \tag{3}$$

The image pixel at time $\tau$ is usually taken to be $|\emptyset_{ss}(\tau)|^2$, which resembles a $[(\sin\chi)/\chi]^2$ function with high -13.2 db sidelobes. The autocorrelation (2) can of course be performed in the frequency domain. Let s(n) represent a sampled version of s(t) and let

$$\begin{aligned} S(\omega) &= \text{discrete fourier transform of } s(n) \\ &= \mathrm{DFT}(s) \end{aligned}$$

so that

$$\begin{aligned} s(n) &= \text{inverse discrete fourier transform of } S(\omega) \\ &= \mathrm{IDFT}(S) \end{aligned}$$

Then if W is a weighting function,

$$\tilde{\phi}_{ss}(\tau) = \text{IDFT}\,(S \cdot S^* \cdot W) \tag{4}$$

is the discrete response to the weighted "matched" filtering. For properly designed $W(\omega)$, the output

$$|\tilde{\phi}_{ss}(\tau)|^2$$

will have peak sidelobe ratios of -35 db or lower. However the impulse response mainlobe will broaden and the peak amplitude will decrease. The choice of an optimal weighting function depends on several factors, such as pulse widening and far sidelobe falloff rate [6]. This problem does not concern us here; we shall use Hamming weighting throughout, with H = 0.08:

$$W(\omega) = \begin{cases} 0.08 + 0.92 \cos^2\left[\pi\left(\dfrac{\omega - \omega_0}{\Delta\omega}\right)\right] & \text{for } |\omega - \omega_0| < \Delta\omega/2 \\ 0 & \text{for } |\omega - \omega_0| > \Delta\omega/2 \end{cases} \tag{5}$$

where $\omega_0$ = spectrum center frequency

$\Delta\omega$ = spectrum bandwidth

This form of weighting ideally gives -42.8 db peak sidelobe ratios and mainlobe null-to-null broadening ratios of 2.0. However, these results can not be achieved with a sampled, quantized signal of finite time bandwidth product. In addition, the presence of an azimuth antenna pattern further reduces the effectiveness of the weighting function, as does quadratic phase error.

This paper reports the results of simulations conducted to determine the relationship between weighting effectiveness on the one hand and quantization, quadratic phase error, time bandwidth product and antenna pattern on the other. The effect of output undersampling has been previously demonstrated in [1].

Section II describes the assumptions and parameters of the simulations. Included is a breakdown of the radar system parameters, chosen for their similarity to the SEASAT radar parameters [5]. Section III describes the quantization algorithm and tabulates the results of varying the quantization precision. Section IV gives the method used for varying the time-bandwidth product, and shows how the TBP affects the weighting effectiveness. Section V reports the results of superimposing an antenna pattern on the simulated return. Finally, Section VI shows the effect of weighting in the presence of quadratic phase errors.

## II. SIMULATION PARAMETERS AND ASSUMPTIONS

Throughout all the simulations referred to in this report, the simulated azimuth radar return was stored in a complex array of 4096 elements. The first 2048 returns represented I and Q samples from a point target. The slant range R to the target at aperture center was 800 km. (The aperture center was located at point 1024 of the array.) The interpulse spacing was assumed to be $\Delta X = 2.0$ meters (except when it was varied to vary the time - bandwidth product; see Section IV). Thus the synthetic aperture length was 2.0 X 2048 meters, or 4.096 km. The radar was assumed to be L-band with a wavelength $\lambda = .235$ meters. It follows that the $k^{th}$ element of the signal array was given by

$$s(k) = \begin{cases} \exp\left\{2\pi j(k - 1024)^2 (\Delta X)^2 / \lambda R\right\} & \text{for } 1 \leq k \leq 2048 \\ 0 & \text{for } \quad k > 2048 \end{cases} \tag{6}$$

The reference signal array was identical to the signal array, except right shifted by 2048 elements. Although the simulated signal and reference are for an azimuth return, the results of this paper are also directly applicable to linear FM range weighting.

The weighting function chosen for all frequency weighting was as in (5). The spatial bandwidth $\Delta\omega$ was determined from the formula

$$\Delta\omega = 2(FMAX) \tag{7}$$

$$FMAX = 2048(\Delta X)/\lambda R$$

This form of weighting provides a practical and realizable approximation to the ideal Dolph-Chebyshev weighting [3].

It is well known that for linear FM waveforms, frequency weighting can be approximated by time weighting across the time reference. In our simulations, a time weighting of the form

$$w(k) = 0.08 + 0.92 \cos^2\left(\frac{\pi(k - 3072)}{2048}\right) \quad (8)$$

$$\text{for } 2049 \le k \le 4096$$

was applied for all time weighting simulations. However, if it is desired to implement a time weighting $w_1(k)$ which produces exactly the same effect as the given frequency weighting $W(\omega)$, then $w_1$ may be obtained from the relations

$$DFT(s \cdot w_1) = S \cdot W \quad (9)$$

$$w_1 = \frac{IDFT(S \cdot W)}{s} \quad (10)$$

where all algebraic operations are performed pointwise. Of course $w_1(k)$ in (10) is taken as 0 for k outside the reference function time duration. Note that $w_1$ and W are not Fourier transform pairs. Time domain weighting is useful where time domain processing is to be implemented, since there is then no need to take the forward Fourier transform of the signal. Furthermore, as will be seen, the effect of a non-ideal finite time-bandwidth product linear FM waveform is such that in some instances, a time domain weighting achieves better results than the corresponding frequency domain weighting.

The block diagram of the simulation for frequency weighting is shown in Figure 1. As seen from the figure, a forward FFT is taken on both the signal and reference arrays, and the two FFT's are then multiplied pointwise. Also, a weighting function is then multiplied pointwise with the product of the two FFT's. Next, an inverse FFT is taken, and the result is pointwise complex absolute value squared. It is this squared output (intensity) that is used to compute all parameters in the tables in the following sections. The corresponding block diagram for time weighting is shown in Figure 2.

The parameters for all tables in this paper are defined as follows:

3 db width = number of array points within the mainlobe which are $\geq$ -3 db below the center peak

0 - 0 width = number of array points between the first two relative minima

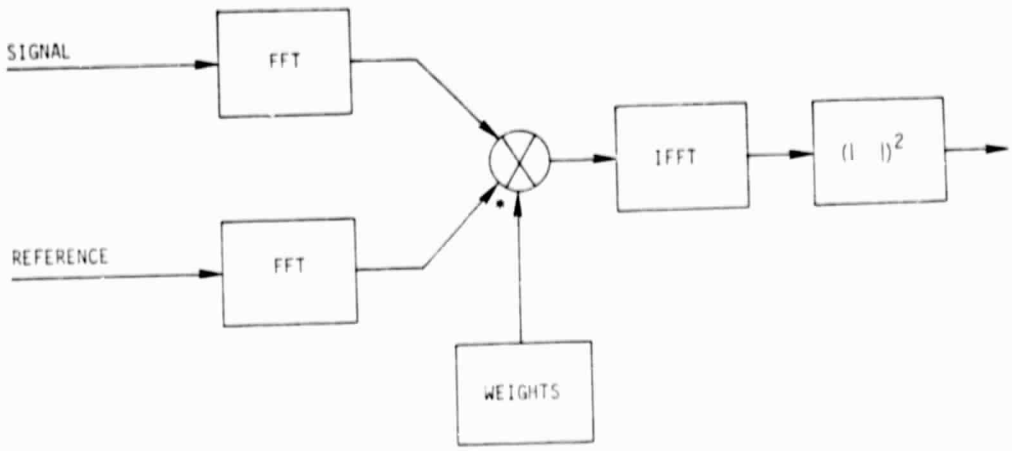


Fig. 1. Frequency Weighting Block Diagram

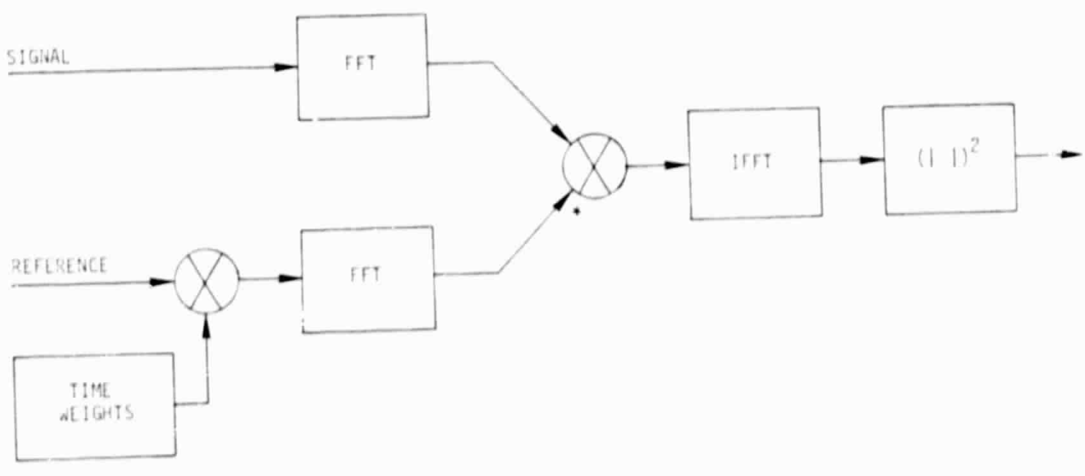


Fig. 2. Time Domain Weighting

$$PSL = \text{peak sidelobe ratio}$$

$$= -10 \log_{10} (\text{peak intensity/highest sidelobe intensity})$$

$$ISLR = \text{integrated sidelobe ratio}$$

$$= 10 \log_{10} (E_o/E_i)$$

where

$$E_i = \text{total energy between first two nulls}$$

$$E_o = \text{total energy outside first two nulls}$$

$$SL = \text{signal loss}$$

$$= 10 \log_{10} (E_p/E_t)$$

where

$$E_p = \text{energy in peak (center) array point}$$

$$E_t = \text{total energy in the array}$$

Note that the SL gives a normalized estimate of the peak response energy. The normalization was necessary since quantization, weighting, etc. all have a scaling effect on the output array.

$$TBP = \text{time bandwidth product (see equation (7))} \qquad (11)$$
$$= (2048 \, \Delta X) \, (\Delta \omega)$$

## III. EFFECT OF QUANTIZATION

The quantization algorithm used in all cases was uniform, bipolar and even. Since the algorithm was not normalized, it was necessary to scale the output for SL measurements as indicated in Section II. Whenever quantization was not performed, 32 bits floating point arithmetic was used.

Several quantization options were examined. The first was frequency weighting with time quantization. The block diagram is shown in Figure 3. Basically, the signal and reference are quantized before the FFT. The frequency weighting function is still 32 bits. The results are summarized in Table 1. Notice that the effect of reference quantization is essentially negligible down to 4 bits.

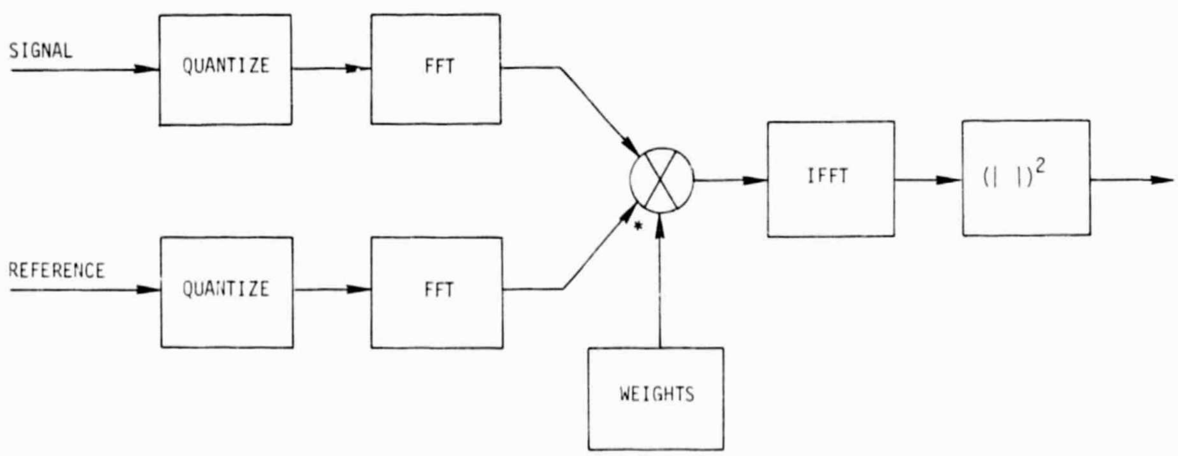


Fig. 3. Frequency Weighting With Time Quantization

Table 1. Frequency Weighting With Time Quantization ($\cos^2$ + Pedestal Weighting) (TBP = 178)

| QUANTIZATION, BITS | | | | | | |
|---|---|---|---|---|---|---|
| SIGNAL | REFERENCE | 3db WIDTH | 0-0 WIDTH | PSL | ISLR | SL |
| 32 | 32 | 17 | 47 | -40 | -24 | -9 |
| 32 | 6 | 17 | 47 | -40 | -24 | -9 |
| 32 | 4 | 17 | 47 | -39 | -24 | -9 |
| 32 | 2 | 15 | 45 | -33 | -19 | -9 |
| 6 | 4 | 15 | 47 | -39 | -24 | -9 |
| 4 | 4 | 17 | 45 | -38 | -24 | -9 |

It was next desired to perform time weighting, so that the signal, reference, time weights and weighted reference could all be quantized. The first step was to select an "optimal" form of time weighting. The different time weightings considered are shown in Table 2, along with their effects. "Ideal" weighting was as in equation (10). Ideal "complex absolute value" weighting was $|w|$, where w again was as in equation (10). Finally, the "$\cos^2$ + pedestal" was as in equation (8), and gave the best results. Therefore it was selected for the time weighting quantization simulations.

Table 2. Time Domain Weighting Results 32 Bit Quantization (TBP = 178)

| TIME WEIGHTING | 3db WIDTH | 0-0 WIDTH | PSL | ISLR | SL |
|---|---|---|---|---|---|
| IDEAL | 17 | 17 | -41 | -26 | -9 |
| IDEAL, COMPLEX ABSOLUTE VALUE | 17 | 47 | -42 | -26 | -9 |
| $COS^2$ + PEDESTAL LENGTH 2048 | 17 | 47 | -42 | -27 | -9 |

The block diagram for these last simulations is shown in Figure 4. Note that effectively two stages of quantization are applied to the reference and time weights. The results are shown in Table 3 for a number of quantization combinations.

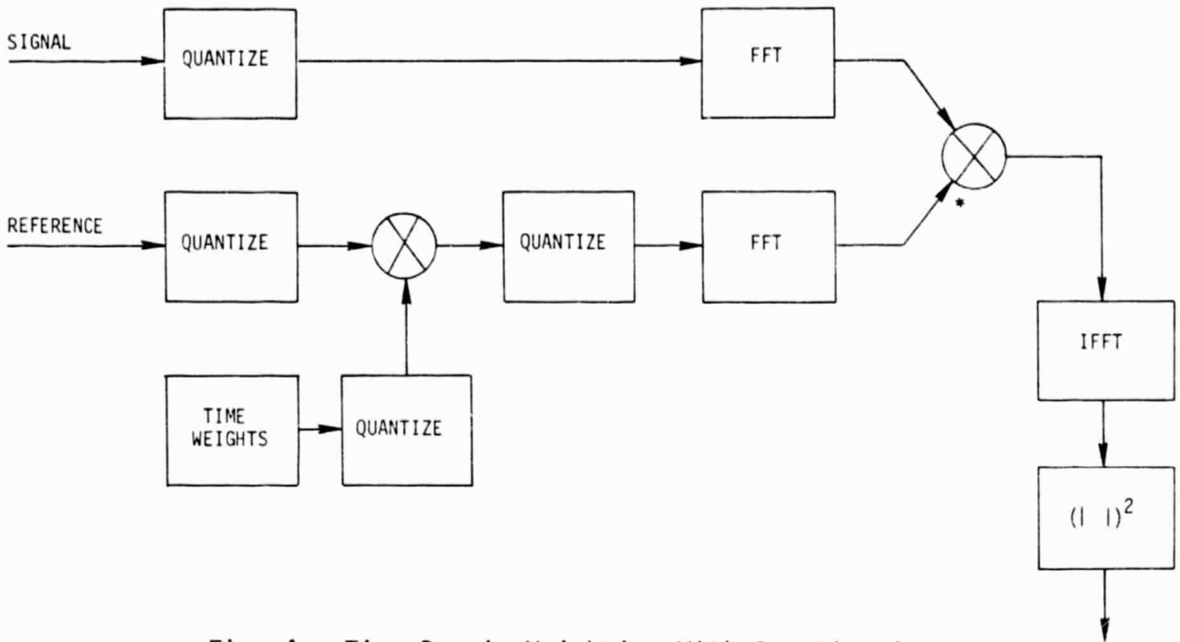


Fig. 4. Time Domain Weighting With Quantization

Table 3. Quantized Time Weighting (TBP = 178)

| QUANTIZATION | | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| SIGNAL | REFERENCE | WEIGHTS | WEIGHTS X REFERENCE | 3db WIDTH | 0-0 WIDTH | PSL | ISLR | SL |
| 32 | 32 | 32 | 32 | 17 | 47 | -42 | -27 | -9 |
| 32 | 8 | 8 | 4 | 17 | 49 | -37 | -26 | -9 |
| 8 | 8 | 8 | 4 | 17 | 49 | -37 | -26 | -9 |
| 6 | 8 | 8 | 4 | 15 | 49 | -37 | -26 | -9 |
| 4 | 8 | 8 | 4 | 17 | 49 | -37 | -25 | -9 |

## IV. TIME BANDWIDTH PRODUCT

In all simulations described thus far, the time-bandwidth product of the signal was, (from Section II)

$$TBP = 178$$

Ideally, the $\cos^2$ + ped frequency weighting should give -42 db for the PSL and -35 db or better for an ISLR with mainlobe broadening ratios of 2.0 [2]. However, such calculations assume a continuous signal with infinite TBP. For our purposes, a signal with infinite TBP was simulated by using a rectangular spectrum and applying frequency weighting. The other (finite) TBP cases were obtained by varying $\Delta X$ in (11) and (7). Therefore the infinite TBP data in Table 4 are not to be compared with the finite TBP data, since the mainlobe widths are not proportional. However, the numbers for PSL, ISLR and SL will serve as a useful reference for comparison. In Table 4 the results with and without weighting are shown along with the results for several different TBP's between 20 and 800. The Table shows that TBP's of over 150 are probably required for effective azimuth weighting.

Table 4. Frequency Weighting vs. TBP (32 Bits Quantization)

| WEIGHTING | TBP | 3 db WIDTH | 0 - 0 WIDTH | PSL | ISLR | SL |
|---|---|---|---|---|---|---|
| None | 22 | 95 | 197 | -13 | -10 | -17 |
| None | 87 | 23 | 49 | -13 | -10 | -11 |
| None | 178 | 11 | 25 | -13 | -10 | - 8 |
| None | 350 | 5 | 13 | -13 | -10 | - 5 |
| None | 714 | 3 | 7 | -13 | -11 | - 3 |
| $\cos^2$ + ped | 22 | 137 | 345 | -24 | -16 | -18 |
| $\cos^2$ + ped | 87 | 33 | 93 | -36 | -21 | -12 |
| $\cos^2$ + ped | 178 | 17 | 47 | -40 | -24 | - 9 |
| $\cos^2$ + ped | 350 | 9 | 25 | -41 | -27 | - 7 |
| $\cos^2$ + ped | 714 | 5 | 15 | -42 | -30 | - 4 |
| $\cos^2$ + ped* | $\infty$ | 47 | 129 | -42 | -36 | -13 |
| None* | $\infty$ | 29 | 65 | -13 | -10 | -12 |

* These entries are not comparable to the others in the table. Please refer to page 11 to see how they were computed.

The reason for the variation in weighting performance is of course due to the non-rectangular power spectrum of small bandwidth linear FM signals. See Figure 5.

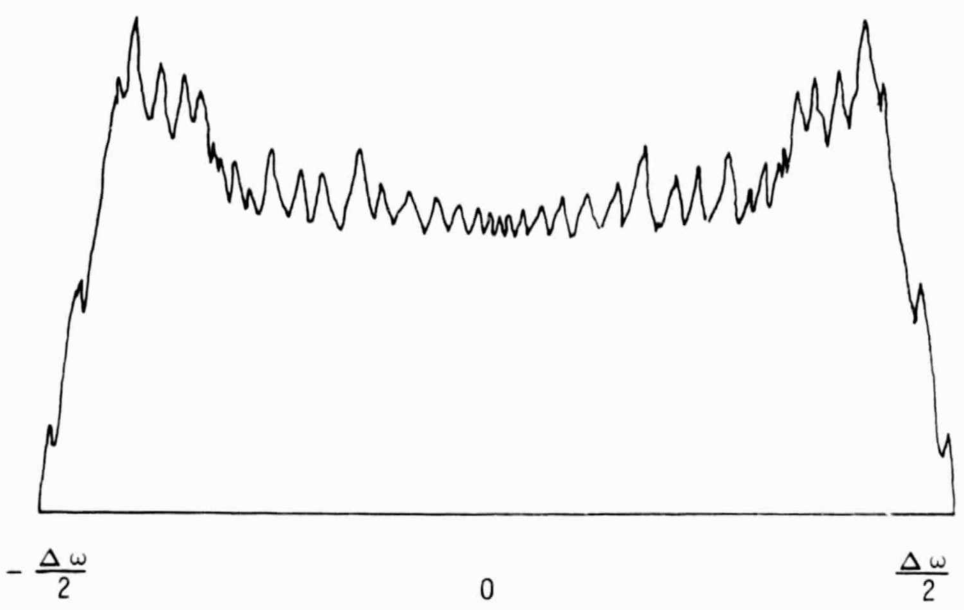


Figure 5. Linear FM Spectrum For Small TBP

## V. THE ANTENNA PATTERN

In weighting the SAR azimuth reference function, it must be remembered that the received signal has already been time "weighted" by the antenna pattern. In the one-look case therefore, if additional weighting is applied to the reference function, the total weighting effect becomes quite pronounced, with significant mainlobe broadening and SL drop.

However, assuming that, say, four looks are to be taken, the antenna pattern no longer resembles a symmetric weighting function within each look. The question then is: What is the combined effect due to weighting and the antenna pattern? It is of course possible to design weighting that compensates perfectly for an antenna pattern as follows. Let $a(k)$ represent the antenna pattern weight to be applied to the $k^{th}$ signal return $s(k)$. Let the reference be $r(k) = s^*(k)$. Then the received signal is

$$\tilde{s}(k) = s(k)a(k) \tag{12}$$

Suppose we have a frequency weighting $W(\omega)$ that works well on $s(k)$ when s has no antenna pattern. Then we would want the adjusted weighting $W_1(\omega)$ to satisfy

$$W_1(\omega)\tilde{S}(\omega) = W(\omega)S(\omega) \tag{13}$$

or

$$W_1(\omega) = \frac{W(\omega) \cdot S(\omega)}{\tilde{S}(\omega)} \tag{14}$$

If it is now desired to construct a time weighting $w_2(k)$ to produce the same effect as $W_1(\omega)$ in (14), then of course $w_2$ is given by

$$\mathrm{DFT}(w_2 \cdot \tilde{s}) = W(\omega)S(\omega) \tag{15}$$

or

$$w_2 = \frac{\mathrm{IDFT}(W \cdot S)}{\tilde{s}} \tag{16}$$

Alternately, $w_2$ may be obtained from (13) by

$$DFT(w_2 \cdot \tilde{s}) = W_1(\omega)\tilde{S}(\omega) \tag{17}$$

so that also

$$w_2 = \frac{IDFT(W_1 \cdot \tilde{S})}{\tilde{s}} \tag{18}$$

Equation (14) gives the frequency weighting; equations (16) and (18) give the time weighting. As in (9) and (10), all algebraic operations are performed pointwise; and again, $w_2$ and $W_1$ are not transform pairs.

Although equations (14), (16) and (18) solve the antenna pattern problem, the simulations reported here indicate that such "fine-tuned" weighting functions may not be necessary for multilook radars, as will be shown in Table 5. In the Table, the antenna pattern was taken to be of the form $\left(\frac{\sin X}{X}\right)^2$ with about 80% of the null-to-null width of the pattern utilized for the SAR azimuth processing. (Note that no antenna pattern sidelobes were included in the simulations. The results might be different if antenna pattern sidelobes (foldover) were to be included. It is, however, extremely difficult to simulate the effect of antenna pattern sidelobes on a point target response). Because of symmetry, only the first two looks have been tabulated. Quantization was fixed at 32 bits, and time weighting (Eq. (8)) was applied. The time-bandwidth product (per look) was fixed at 178. In fact, the bandwidth is constant for all table entries, so that the 1-look resolution without weighting is the same as the 4-look resolution without weighting. (See rows 1 and 5 of the Table). The 4-look aperture here would thus be required to be 4 times as long as the 1-look aperture. When multiple looks are taken, each look utilizes only a fraction of the antenna pattern. When only one look is taken, it would presumably utilize almost all of the mainlobe of the antenna pattern. Thus the presence of the antenna pattern should induce a weighting effect when only one look is taken, but should not induce a weighting effect when multiple looks are taken. This conclusion is verified by observing that in the single look mode, rows 1 and 3 of the table show that the presence of the full antenna pattern produces a strong weighting effect in terms of mainlobe broadening and ISLR improvement. In the multilook mode, rows 5 and 7 may be compared to show that the presence of the partial per-look antenna pattern produces only a negligible weighting effect. (The ISLR's are only 1 db apart, and the mainlobes have the same width). The same observation applies to a comparison between rows 6 and 8, where weighting has been applied with and without an antenna pattern. Note from rows two and four of the table that the single look antenna pattern has a strong weighting effect even when system weighting is applied separately.

Table 5. Weighting With An Antenna Pattern

| ANTENNA PATTERN | LOOK NUMBER | WEIGHTS | 3 db WIDTH | 0 - 0 WIDTH | PSL | ISLR | SL |
|---|---|---|---|---|---|---|---|
| No | 1 of 1 | No | 11 | 25 | -13 | -10 | - 8 |
| No | 1 of 1 | Yes | 17 | 47 | -42 | -27 | - 9 |
| Yes | 1 of 1 | No | 17 | 43 | -35 | -26 | - 9 |
| Yes | 1 of 1 | Yes | 19 | 71 | -46 | -57 | -10 |
| No | 1 of 4 | No | 11 | 25 | -13 | -10 | - 8 |
| No | 1 of 4 | Yes | 17 | 47 | -42 | -27 | - 9 |
| Yes | 1 of 4 | No | 11 | 25 | -14 | -11 | - 8 |
| Yes | 1 of 4 | Yes | 17 | 49 | -43 | -27 | - 9 |
| Yes | 2 of 4 | No | 11 | 25 | -14 | -11 | - 8 |
| Yes | 2 of 4 | Yes | 17 | 49 | -43 | -28 | - 9 |

## VI. EFFECT OF QUADRATIC PHASE ERRORS

Quadratic phase mismatch between the radar signal and the reference function is a further factor to be considered in examining the effect of weighting. This mismatch is due to inaccurate knowledge of the radar sensor position and altitude parameters over the length of the synthetic aperture. If we suppose that the ideal matched filter azimuth impulse response closely resembles a sin X/X function, then quadratic phase error will tend to broaden and deform the mainlobe to varying degrees.

Using the notation of equation (2), let us assume that the (non-ideal) reference function $\tilde{s}$ has the (incorrect) quadratic parameter b'. Then equation (2) is rewritten as

$$\phi_{s\tilde{s}}(\tau) = \int_{-T/2}^{T/2-\tau} \exp\left\{j\,\frac{b}{2}\,t^2\right\}\,\exp\left\{-\,j\,\frac{b'}{2}\,(t+\tau)^2\right\}dt \qquad (19)$$

In the case where no weighting is applied, it is shown in [2] and [4] that quadratic phase errors of $\pi/4$ at the aperture edges lead to 3 db mainlobe broadening of less than 15%, and null-to-null mainlobe broadening of less than 10%. Since the impulse response is deformed in the presence of quadratic phase errors, an alternate definition of resolution is needed. Define 3 db resolution as the distance from the peak to the point at which 71.7% of the area under the curve has been exhausted. Similarly, define null-to-null resolution as the distance from the peak to the point at which 90.3% of the area under the curve has been exhausted. These definitions reduce to the standard definitions in the case where there is no quadratic phase error or weighting [4]; i.e. the 3 db point of an unweighted response encompasses 71.7% of the area under the curve.

To simulate azimuth quadratic phase error, we used the fact that the quadratic phase characteristic at the aperture edge is given by

$$\phi = \frac{2\pi L^2}{\lambda R} \qquad (20)$$

where

$2L$ = total aperture length

$\lambda$ = wavelength

$R$ = slant range at aperture center

Thus if we assume that the reference function incorporates an erroneous slant range factor $R_{err}$, we obtain

$$\phi_{err} = \frac{2\pi L^2}{\lambda R_{err}} \qquad (21)$$

as the (erroneous) quadratic phase characteristic of the reference function at the aperture edge. Therefore

$$\Delta\phi = \phi - \phi_{err} = \frac{2\pi L^2}{\lambda}\left(\frac{1}{R} - \frac{1}{R_{err}}\right) \qquad (22)$$

is the quadratic phase error at aperture edge due to a range mismatch of

$$\Delta R = R - R_{err}$$

Equations (20), (21) and (22) were used to simulate quadratic phase errors in the presence of weighting. The weighting function was as in (5). The results are shown in Table 6, with the unweighted, $\Delta\phi = 0$ parameters all normalized. The second row is from [4]. The reason for the apparent "shrinkage" in row 4 column 2 is the switch to the area criterion definition of resolution. Note that weighting tends to attenuate the degradation due to quadratic phase errors, as predicted in [2]. (Compare rows one, two and five to verify this fact.)

Table 6. Effect of Quadratic Phase Errors in the Presence of Weighting

| CASE | 3 db WIDTH | NULL-TO-NULL WIDTH | PSL | ISLR db | SL INDICATOR db |
|---|---|---|---|---|---|
| UNWEIGHTED $\Delta\varphi = 0$ | 1 | 1 | -13 | -10 | 0 db |
| UNWEIGHTED $\Delta\phi = 180°$ | 3.0 | 2.3 | - | - | - |
| WEIGHTED* $\Delta\phi = 0°$ | 1.2 | 1.4 | -24 | -18 | -1 |
| WEIGHTED $\Delta\phi = 90°$ | 1.4 | 1.033 | - | - | -1 |
| WEIGHTED $\Delta\phi = 180°$ | 2.2 | 1.57 | - | - | -3 |
| WEIGHTED $\Delta\phi = 270°$ | 3.07 | 2.2 | - | - | -6 |
| WEIGHTED $\Delta\phi = 360°$ | 4.2 | 2.83 | - | - | -7 |

* In this case, the actual 3 db drop and zero crossing were located. If the area criterion is used, we find the 3 db width is 1.133 and the null-to-null width is 0.767.

## VII. SUMMARY

The time-bandwidth product of a digitally processed SAR signal has a dramatic impact on the effectiveness of sidelobe reduction weighting. For a one-look radar, the antenna pattern may eliminate the need for weighting. For a multi-look radar (with $\geq$ 4 looks) weighting may be implemented without regard for the antenna pattern. Quantization to 4 bits provides minimal PSL and ISLR degradation, while weighting effectiveness is only slightly reduced by quadratic phase error.

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[6] "Effect of Quantization on Integrated Sidelobe Ratios VOIR Action Item #9," JPL Interoffice Memorandum 3341-193-78 from H. Zebker to D. Rodgers, March 7, 1978. (JPL internal document.)