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Measuring channel balance in multi-channel radar receivers

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ABSTRACT

Radar receivers with multiple receive channels generally strive to make the receive channels as ideal as possible, and as alike as possible. This is done via prudent hardware design, and system calibration. Towards that end, we require a quality metric for ascertaining the goodness of a radar channel, and its match to sibling channels. We propose a relevant and useable metric to do just that.

Keywords: radar; multi-channel; balance; metric

1 INTRODUCTION

A simple radar system transmits from a single source, and receives with a single antenna, and further processes with simple algorithms a single echo signal characteristic. Such a receiver exhibits a single receiver channel. Additional information from a diversity of received signal characteristics often allows target discrimination with increased reliability, sensitivity, precision, and/or accuracy. Such systems will generally process two or more distinct signal characteristics, each with its own processing channel. Examples of multi-channel radar receivers include Interferometric Synthetic Aperture Radar (IFSAR, InSAR), Ground/Dismount Moving Target Indicator (GMTI/DMTI), tracking radars, direction-finding modes, polarimetric radars, and quadrature demodulation channels.

In all cases, the distinct radar channels seek to process specific inputs so that differences in the inputs can be analyzed and assessed. That is, the purpose of the different channels is to help assess the different "inputs," with differences in the channels themselves made irrelevant to the final comparison. We want our final answer to be about the input signals, and not confused by variations or perturbations of the signal path within the radar. This necessitates eliminating differences in the processing channels, or at the very least identifying and understanding them.

While we desire that multiple channels all exhibit ideal characteristics with respect to processing or in the transmission of signals, alas we must allow that real systems will not be perfect. However, with multiple channels we are often more interested in "differences" of their signals, implying that "relative" channel performance properties may have different requirements, even stricter requirements, than the individual channels with respect to ideal. That is, channel "balance" requirements are often different, and perhaps more strict, than their individual allowable non-idealness.

The authors have repeatedly observed that well-calibrated and balanced channels often make the difference between success and failure of a radar application/mission. Clearly, to fix something, you first must be able to measure it. Otherwise you can't know that it even needs fixing, let alone that it was ever fixed thereafter. Consequently, in this report we examine channel balance measures, and specific balance metrics.

This paper is an abridged version of a more comprehensive report on this topic by the authors.[1]

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2 DETAILED DISCUSSION

2.1 Non-Ideal Channel

We define a general channel transfer function, or set of channel transfer functions as

$$H_k(f) = A_k(f) e^{j\Phi_k(f)} \operatorname{rect}\left(\frac{f - f_k}{B_k}\right), \tag{1}$$

where the constituent parameters and functions are k = index denoting a particular channel, $A_k(f)$ = positive real-valued amplitude function across the passband for channel k, $\Phi_k(f)$ = real-valued phase function across the passband for channel k, f_k = nominal center reference frequency for the passband for channel k, B_k = nominal bandwidth of the passband for channel k, and $T_k(f)$ = nominal group delay of a signal through the channel for channel k. For completeness, we also identify the corresponding channel impulse response with our designation for a Fourier Transform pair as

$$h_k(t) \Leftrightarrow H_k(f). \tag{2}$$

The rect function in Eq. (1) is defined as

$$\operatorname{rect}(\zeta) = \begin{cases} 1 & |\zeta| < 1/2 \\ 1/2 & |\zeta| = 1/2 \\ 0 & else \end{cases}. \tag{3}$$

We identify an ideal channel as one whose amplitude function is constant, and phase function is linear in frequency *f*. As a practical matter, for transmission channels, we normally expect the amplitude function $A_k(f)$ to be relatively smoothly varying with frequency *f* across the passband. We furthermore normally expect that the unwrapped phase function $\Phi_k(f)$ will also be a relatively smoothly-varying function of frequency *f*, and with a typically strong linear component.

The group delay parameter T_k is not overtly present in the channel function as written in Eq. (1), but may be considered related to the linear phase component in $\Phi_k(f)$; in particular its frequency slope. It will become important later in this paper. For now, we simply identify it as the group delay of the channel and generally a function of frequency, that is

$$T_k(f) = -\left(\frac{1}{2\pi}\right)\frac{d}{df}\Phi_k(f). \tag{4}$$

Although mentioned above, we have made the tacit assumption that the phase is "unwrapped," that is, does not contain any 2π jumps. If we need a single number and not a varying function of frequency, then we might calculate the group delay at band center, or perhaps some average value over the band of interest. This might also be related to the location of the peak of the impulse response.

The ideal situation for us is that all channels in a set are alike, and in fact identical to the ideal channel previously discussed. Failing being ideal, we would then desire that all channels at least be alike. The intent is that any differences in channel outputs be attributed to differences in their inputs, and not due to differences in the channels themselves. Sadly, neither of these situations can be reasonably expected with real hardware. Consequently, we need a measure of similarity between channels, including the ideal channel, to gauge non-ideal characteristics.

Finally, we will proceed with the assumption that any measurements of channel properties are high-fidelity, collected in a manner so as to be minimally impacted by measurement noise, or other measurement distortion.

2.2 Channel Similarity Measure

The conventional technique of measuring channel/filter characteristics is with a network analyzer, which measures gain and phase shift over a set of frequencies. That is, it measures the transfer function over a set of sample frequencies. Using 2-port Scattering-parameters (S-parameters), this is the S_{21}, or forward voltage gain, measure. Consequently, we will presume as input to any similarity measure the channel transfer function, and will calculate other measures from these.

We root our similarity measure in the concept of Euclidean distance, or L^2 norm. This is tractable because energy/power is proportional to the square of signal voltage, so minimizing the squared-error is minimizing the energy/power in the error. We furthermore normalize the squared-error measure such that

$$\eta(f) = \frac{|E_{k,l}(f)|^2}{|H_k(f)|^2 + |H_l(f)|^2} = \frac{|H_k(f) - H_l(f)|^2}{|H_k(f)|^2 + |H_l(f)|^2} = \text{normalized error measure.} \tag{5}$$

We are, of course, assuming no mutual nulls in the channel ESDs, at least over the frequency spans of interest. In fact, there is an implicit limitation that this error measure is for a common frequency span. Based on this normalized error, we define a Figure of Merit (FOM), and then expand it as

$$FOM(f) = 1 - \eta(f) = \left[\frac{\left(\sqrt{A_k^2(f) A_l^2(f)} \right)}{\left(\frac{A_k^2(f) + A_l^2(f)}{2} \right)} \right] \cos\left(\Phi_k(f) - \Phi_l(f)\right). \tag{6}$$

We have retained this FOM as a function of frequency. The FOM is real-valued for all relevant frequencies. If both channels match each other perfectly, the normalized error is zero, and the FOM is unity for all relevant frequencies. We also note that the FOM is a product of contributions from

1. A function of the phase differences, and
2. A function of the individual amplitude terms (the individual ESD).

One component of our FOM is due to phase differences, and is contained in the cosine term. Clearly, this FOM component is between −1 and +1, with +1 only possible if the phases are equal.

The other component of our FOM is due to amplitude differences, and is contained in the square brackets. We observe that the numerator is a geometric mean of the ESDs, and the denominator is the arithmetic mean of the ESDs, both at a specified frequency. In general, by the theorem of inequality of arithmetic and geometric means, this ratio is less than or equal to unity, with equality holding only when $A_k^2(f) = A_l^2(f)$, and since amplitude functions are by definition positive, only when $A_k(f) = A_l(f)$. Consequently, this FOM component is between zero and +1.

2.3 Channel Similarity Measure – Small Errors

Consider now the case of the error between two channels being small. We might reasonably expect this from well-designed hardware. Accordingly, with malice aforethought, we define a ratio of channel transfer functions as

$$R_{k,l}(f) = \frac{H_k(f)}{H_l(f)}. \tag{7}$$

The tacit assumption is that $H_l(f)$ exhibits no nulls in the passband. Again, this does not seem to be unreasonable for well-designed hardware. This means we can write one transfer function in terms of the other as

$$H_k(f) = R_{k,l}(f) H_l(f). \tag{8}$$

Using this with Eq. (6) and Eq. (7) allows us to write the Figure of Merit as

$$FOM = \frac{\mathrm{Re}\{R_{k,l}(f)\}}{\mathrm{Re}\{R_{k,l}(f)\} + \frac{1}{2}\left|R_{k,l}(f) - 1\right|^2}, \tag{9}$$

which can for small errors be approximated as

$$FOM \approx 1 - \left(\mathrm{Re}\{R_{k,l}(f)\} - 1\right)^2. \tag{10}$$

The adequacy of this approximation depends on just how small really is the initial imbalance, whether by design or some degree of preliminary calibration. Nevertheless, it gets more accurate the closer we get to balancing the channels of interest. Recall that this FOM reaches its maximum value of one only when $R_{k,l}(f) = 1$, that is, when both channels are identical in both amplitude and phase. For matched filters/channels, we obviously desire that over the passband of interest we have an ideal response, that is

$$R_{k,l}(f) = \mathrm{rect}\left(\frac{f - f_k}{B_k}\right) = \mathrm{rect}\left(\frac{f - f_l}{B_l}\right). \tag{11}$$

The ratio of transfer functions may be written in a manner so that we can separately identify amplitude and phase effects. That is, we may write

$$R_{k,l}(f) = R_{amplitude,k,l}(f) R_{phase,k,l}(f), \tag{12}$$

where we can individually identify

$$\begin{aligned} R_{amplitude,k,l}(f) &= \frac{A_k(f)}{A_l(f)} \mathrm{rect}\left(\frac{f - f_k}{B_k}\right), \text{ and} \\ R_{phase,k,l}(f) &= e^{j(\Phi_k(f) - \Phi_l(f))} \mathrm{rect}\left(\frac{f - f_k}{B_k}\right), \end{aligned} \tag{13}$$

where we have made the implicit assumption of identical passbands. Separating amplitude and phase contributions can be useful when specifying channel matching requirements when we have different levels of control for mitigating the effects of imbalance with respect to phase versus amplitude. Important parameters in channel balance measurement are the relative phase difference and the relative group delay between channels. We now define corresponding measures as

$$\begin{aligned} \phi_{k,l} &= \Phi_k(f_k) - \Phi_l(f_k) = \text{nominal relative phase difference, and} \\ \tau_{k,l} &= \left\langle -\frac{1}{2\pi}\frac{d}{df}\left(\Phi_k(f) - \Phi_l(f)\right)\right\rangle = \text{mean relative group delay,} \end{aligned} \tag{14}$$

where $\langle\ \rangle$ denotes mean value, with the tacit assumption that the phase difference is "unwrapped," that is, does not contain any 2π jumps.

2.4 Comparing Channels to Ideal

Recalling the definition of an ideal channel allows us explicitly to write the ratio of transfer functions when comparing to the ideal channel as

$$R_{k,ideal}(f) = \frac{e^{j2\pi T_{ideal} f}}{C_{ideal}} H_k(f), \tag{15}$$

where C_{ideal} is the ideal channel gain, and T_{ideal} is the ideal channel group delay.

Note that the shape of the ratio is determined by the channel filter function $H_k(f)$. The ratio impulse response is merely a scaled and shifted version of $h_k(t)$, with the amounts of scale and/or shift having been predetermined to adjust for whatever is "ideal." The nature of $h_k(t)$ is such that we would expect a relatively strong peak shifted in time from $t = 0$. For an ideal channel it would be a sinc function. For a non-ideal channel we expect a perturbation from the sinc function due to real hardware being composed of non-ideal components. The most obvious effects in $h_k(t)$ will generally be observed in

1. A broadening of the mainlobe,
2. An undesirable scaling in mainlobe amplitude,
3. A shifting in time of the mainlobe, and
4. An amplification of the sidelobes.

That these features are readily observable in $h_k(t)$, and largely distinct from each other, makes the impulse response particularly attractive for assessing channel/filter performance. We stipulate that low-level sidelobe structure might (read that as "probably will") require data tapering (window functions) be applied prior to impulse response calculations. There are a plethora of window taper functions from which to choose.[2]

In any case, the impulse response measure integrates many of the aspects of the transfer function, obviating requirements for, say, quadratic phase error, cubic phase error, random phase errors, and ripple in both amplitude and phase across the passband.

2.5 Comparing Channels to Each Other

We recall the ratio of transfer functions with components as expressed in Eq. (12), Eq. (13), and Eq. (14). From these, we may calculate the corresponding impulse response as

$$r_{k,l}(t) \Leftrightarrow R_{k,l}(f). \tag{16}$$

Our measures will be with respect to the impulse response $r_{k,l}(t)$. We again stipulate that low-level sidelobe structure might (read that as "probably will") require data tapering (window functions) be applied prior to impulse response calculations.

2.6 Relationship to Coherence

We state without elaboration that our FOM is closely related to coherence, noting that coherence is typically an integrated measure, over some bandwidth, rather than a function over frequency. An excellent discussion of coherence for Synthetic Aperture Radar (SAR) images is given by Bickel,[3] and using coherence as a measure of goodness for SAR image compression is offered by Bickel and Doerry.[4]

2.7 Ideal Nonlinear Channels

We now abandon the notion of a linear channel, in the sense that it is common for receiver channels to employ frequency conversion; translation to/from intermediate frequency bands or basebands, for easier data processing. In such systems, the passband into the receiver channel is not the same passband out of the receiver channel; frequency translations and sometimes frequency multiplications occur. Strictly speaking, transfer functions defined based on LTI systems no longer apply. However, if we employ a common input signal, then all we need to do is compare the outputs of the two channels to each other to evaluate their balance, regardless of the mixing, multiplying, or other circuit functions present along the way in the individual channels.

2.8 Secondary Issues

Heretofore we have focused on channel gain functions to signals that are input to them. However, other characteristics of multi-channel systems can also have important impact on overall system performance. We mention here some additional issues to be considered when comparing multiple channels with each other.

2.8.1 Noise Floor

Optimal processing across channels assumes known, and generally equal, background noise levels in the various channels. Ideally, the noise floor is at theoretically minimum levels, unless specifically designed to be otherwise.[5] If subsequent processing was to assume equal noise statistics on all relevant channels, when in fact we don't have that, the result may be suboptimal processing results, and perhaps unacceptable biases in relative calculations. Consequently, we desire equal noise spectral statistics, in both shape and level, between all channels. This is especially true for broadband background noise. While noise spectra should exhibit equal statistics, preferably broadband and white, the noise should nevertheless be uncorrelated between channels, lest the noise bias actual signal measurements.

In addition to broadband background noise, a channel may exhibit decidedly non-white spurious content. These often manifest as interfering signals, affecting the fidelity of renderings of the radar data, e.g. range/Doppler images; capable of being detected as false alarms.

2.8.2 Nonlinear Distortions

If any individual channel exhibits nonlinear gain characteristics, then the result is often the generation of harmonic content and mixing products, i.e. spurious signal content of a multiplicative nature. These are, of course, undesirable. A principal source of such nonlinear gain characteristics is an Analog to Digital Converter (ADC).[6,7] We further note that the relative spurious signal levels are often dependent on input signal levels in a very nonlinear manner; they scale in unpredictable ways. Consequently, signal dynamic range is an important factor. Effects of nonlinear distortions are especially important for radar modes that exhibit large integration gains but still wish to detect signals down near the noise level, such as with high-performance GMTI radar systems.[8]

2.8.3 Crosstalk

The typical, and ideal, assumption is that each channel processes only its own desired input. That is, the output of channel k is not influenced by the signal in channel l. Of course, this is ideal, and we are never quite so fortunate. The ability of a signal in one channel to leak, or couple, to another channel is termed "cross-talk." As such, it can be considered a noise source that is highly correlated with a signal in another channel. Clearly, if signals in separate channels are expected to be orthogonal, cross-talk obviously diminishes that orthogonality.

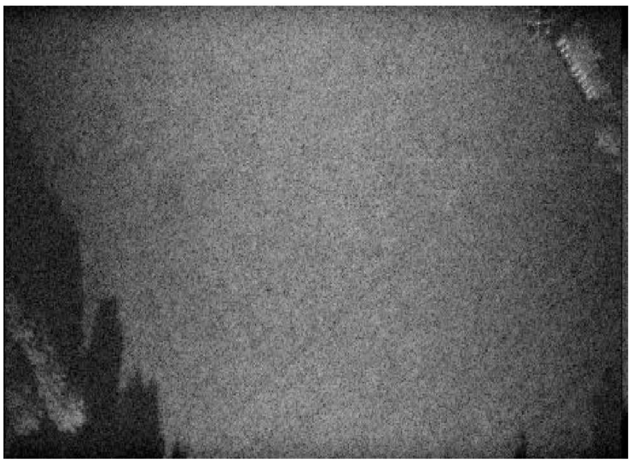
3 EXAMPLES OF IMBALANCE EFFECTS

We now examine the effects of channel imbalance qualitatively on several applications of multiple channels.

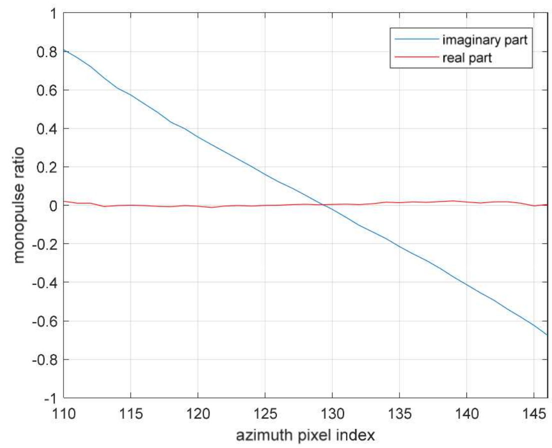
3.1 Monopulse Antenna

Monopulse antennas are used extensively in tracking radars, and for Direction of Arrival (DOA) measurements. An excellent text on these antennas is by Sherman and Barton.[9] Recall that monopulse channels may originate from separate antenna beam feeds, but are typically transmogrified to sum and difference channels by circuitry often referred to as a "comparator." Other names include "monopulse-network," "hybrid," and "magic-T."

Figure 1 shows a SAR image from Sandia's FARAD testbed radar image, as well as the monopulse ratio from the monopulse antenna employed by this radar. Of significance is that the real part of the ratio is essentially zero, and the imaginary part is essentially linear with negative slope, crossing zero at approximately pixel index 129.



(a)



(b)

Figure 1. (a) Ka-band SAR image of grass park from SNL FARAD system, and (b) Mean monopulse ratio with well-balanced channels.

In the following sections we illustrate the effects of amplitude and phase errors that are introduced before (pre-comparator) and after (post-comparator) the formation of the sum and difference channels. More information on the general effects of pre-comparator and post-comparator errors are presented in another report.[10] Unless these errors are detected and corrected, they distort the relationship between the monopulse ratio and estimated angle-of-arrival.

Pre-comparator errors

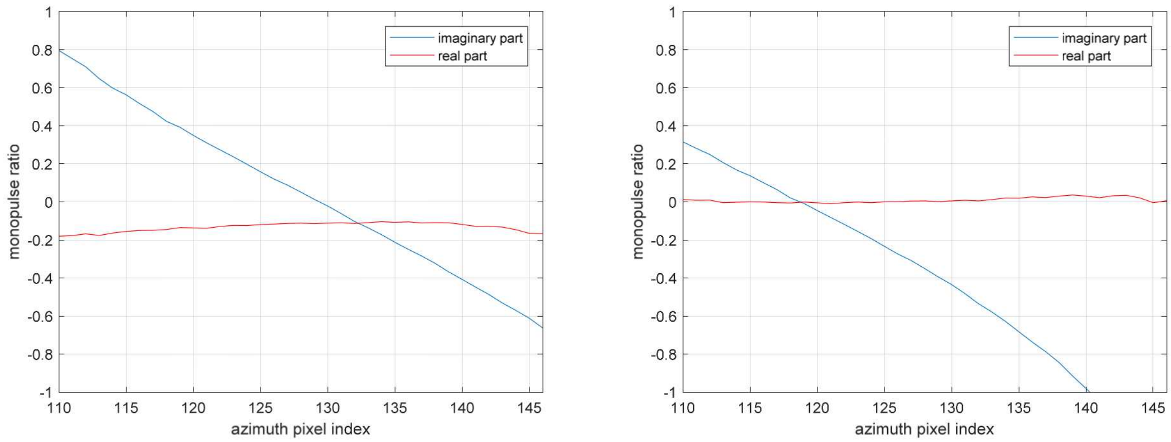
As the name implies, pre-comparator errors are imbalances and error effects that occur before the comparator sums and differences the individual beams. Figure 2(a) shows the error from a pre-comparator amplitude error of about 1 dB. These errors tend to distort the null-depth and location.

A pre-comparator phase error will similarly degrade the null location. Figure 2(b) also shows the error from a pre-comparator phase error of 45º.

Post-comparator errors

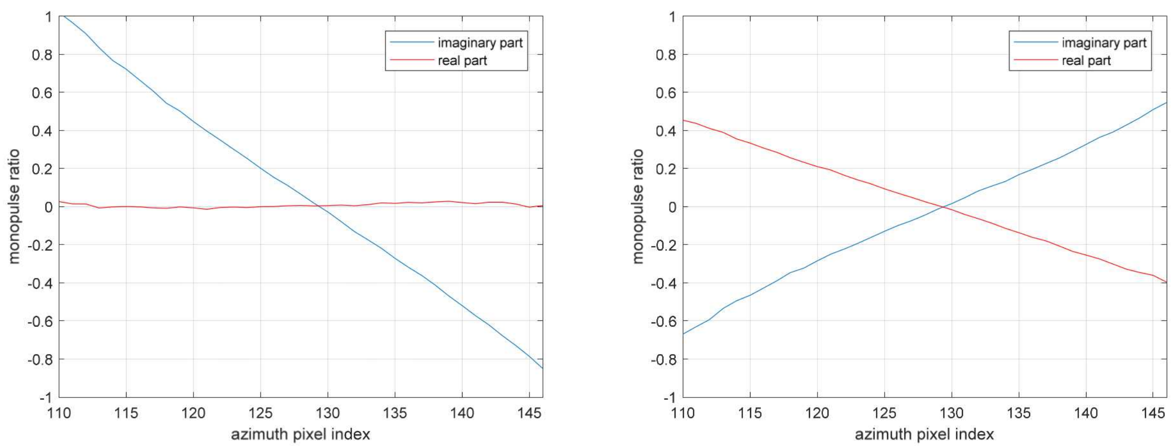
Figure 3(a) shows the error from a post-comparator amplitude error of about 1 dB. A post-comparator amplitude error scales the monopulse ratio which in turn yields a scaling error in the estimated direction-of-arrival.

The post-comparator phase error rotates the monopulse ratio information from the imaginary (from phase monopulse) channel to the real channel which scales the monopulse ratio relative to the direction-of-arrival estimates. Figure 3(b) also shows the error from a post-comparator phase error.



(a) (b)

Figure 2. (a) Mean monopulse ratio with pre-comparator amplitude error, and (b) Mean monopulse ratio with pre-comparator phase error.



(a) (b)

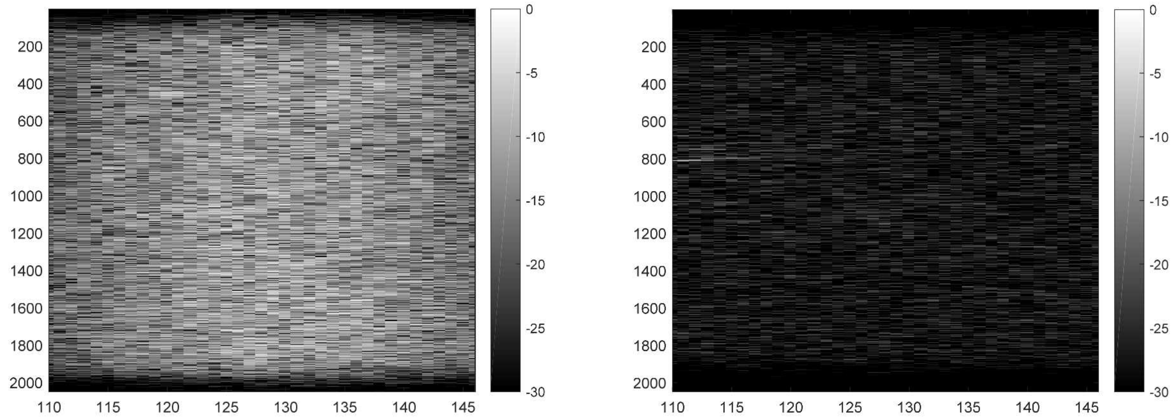
Figure 3. (a) Mean monopulse ratio with post-comparator amplitude error, and (b) Mean monopulse ratio with post-comparator phase error.

3.2 Clutter Cancelling GMTI

In clutter cancellation, uncorrected phase, amplitude, and group delay lead to loss in cancellation performance. Figure 4 shows the short CPI sum image of a region of clutter, where pixel brightness is in units of decibels (dB). Figure 4 also shows the clutter cancelled image of the same area. Errors in cancellation will cause an increase in the residual clutter level which will interfere with target detection. Figure 5 illustrates the effects on clutter cancellation from a 1-dB amplitude error, as well as the effects on clutter cancellation from a 30-degree phase error.

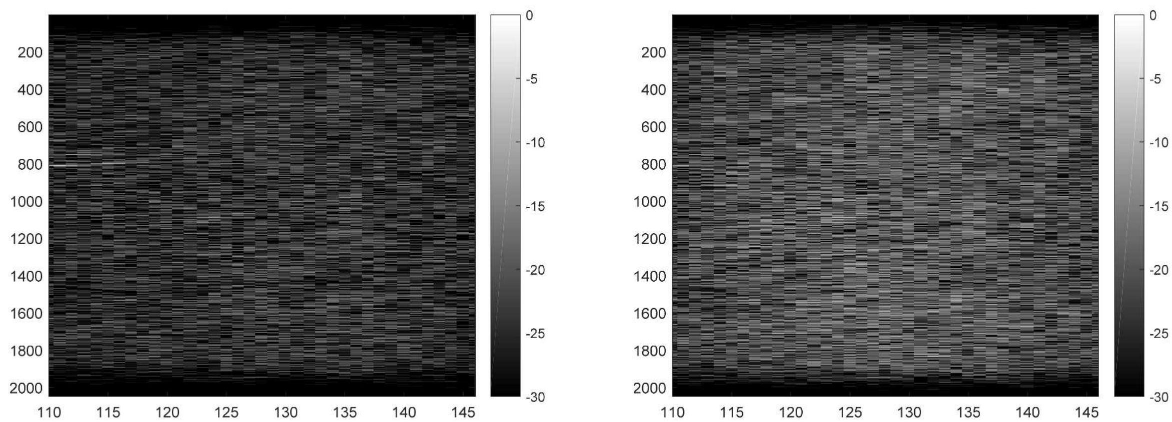
3.3 Interferometric SAR

In Interferometric-SAR (IFSAR, InSAR), calibration errors in phase introduce errors in the estimated terrain. A discussion of height error sources is given in an earlier report.[11] Amplitude calibration errors are less important for IFSAR, but losses in SNR in one of the channels limits coherence, which increases noise in the resulting terrain map. Group delay differences between the channels introduce decorrelation, causing increased noise in the terrain estimate.[4]



(a) (b)

Figure 4. (a) Section of range-Doppler image of uniform clutter, and (b) Reference clutter-cancelled image (mean ~ –26.7 dB)



(a) (b)

Figure 5. (a) Clutter cancellation performance for 1 dB amplitude error (mean ~ –21.7 dB), and (b) Clutter cancellation performance for uncompensated 30-degree phase error (mean ~ –17.8 dB).

3.4 Digital Beamforming (DBF)

In Digital Beamforming (DBF) applications, individual channels generally correspond to signal streams from separate antenna phase centers, or antenna beams. As such, channel errors are indistinguishable from signal modulations. The net results are unintended antenna beam sidelobes and sidelobe characteristics. Intended nulls based on calculations using desired or ideal antenna properties are no longer nulls in practice. This leads to reduced detection probabilities, and increased false alarm probabilities. This is in fact a generalization of the results exemplified in the previous section on clutter cancellation.

A popular technique for processing multiple-channel GMTI data is Space-Time Adaptive Processing (STAP). The intent of STAP processing is essentially to "learn" from the data and use data characteristics to optimize the detection and/or characterization of moving targets. However, the data embody both the effects of the channels as well as the nature of the environment that the radar is interrogating, and data processing is generally unable to separate them without additional information. Often, simplifying assumptions are made in lieu of additional information. However, simply attributing measurements to the environment while ignoring channel differences may lead to erroneous interpretations. Therefore, even in adaptive systems, channel balancing remains important.

3.5 Polarimetric SAR

The desire of channel balance is to maintain polarization orthogonality of the signals in the polarization channels in order to measure target characteristics. As with the previous multiple-channel modes, we want to manage phase, amplitude, and delay differences between the channels. In addition, another important balance correction for multiple-channel radars is the cross-talk measurement and correction. There is more literature on the cross-talk correction for polarimetric radars than for other multiple channel radar system types.[12,13]

3.6 Quadrature Demodulation

Modern high-performance radar systems often operate on the down-converted analytic representation of the received signal energy. The resulting baseband signal is complex-valued, itself represented with In-phase (I) and Quadrature (Q) values that often result from separate channels, or channel segments. To the extent that the I and Q channels are not identical, an imbalance may result, and often does. Imbalance in I/Q channels results in nonorthogonality of their signals, causing leakage of 'image' frequencies into the data, further causing ghosting in radar range-Doppler maps resulting in false targets, often appearing similar to spurious signals.[14]

4 HOW GOOD IS GOOD ENOUGH?

The question now at hand is "How well should receiver channels be balanced, both with respect to ideal, and with respect to each other?" Ultimately, this is an issue of signal purity. Towards this end, we make some rather arbitrary statements below about to what degree channels should be balanced, with the caveat that we do not discourage the reader from developing more reasoned requirements derived from overall system performance requirements.

4.1 Comparing Channels to Ideal

We offer that in the absence of specific derived requirements, any particular radar subsystem, or cascaded subsystems, be correctable to the following.

1. Impulse response mainlobe broadening less than 10 percent from ideal,
2. Impulse response mainlobe scaling be within 1 dB of ideal, for all system gain settings,
3. Group delay be measurable to less than 10 percent of the impulse response mainlobe width, and
4. Impulse response sidelobes less than −40 dBc, with respect to the impulse response mainlobe peak. This might be relaxed somewhat very near the mainlobe peak. This should be measurable with suitable data tapering.
5. Background noise should be white (flat spectrum) at theoretically justified levels; dictated by system noise figure allowances.
6. Any spurious signals, whatever the source, should be at a level below any detection threshold. In the absence of other criteria, spurious levels should be no more than 10 dB above the broadband noise floor, with statistical occurrence consistent with the allowable false-alarm rate.[8]

4.2 Comparing Channels to Each Other

We offer that in the absence of specific derived requirements, any particular radar subsystem, or cascaded subsystems, be correctable to the following.

1. Ratio impulse response mainlobe broadening less than 10 percent from ideal,
2. Ratio impulse response mainlobe scaling be within 1 dB of each other, for all system gain settings,
3. Ratio impulse response mainlobe phase be less than 1 degree at mainlobe peak,
4. Group delay be less than 10 percent of the ratio impulse response mainlobe width, and

5. Ratio impulse response sidelobes be less than –40 dBc, with respect to the mainlobe peak. This might be relaxed somewhat very near the mainlobe peak. This should be measured with suitable data tapering.
6. Broadband noise spectral densities should be within 1 dB of each other for any channel pairs.
7. Crosstalk should meet the spurious signal requirements.
8. We offer that in the absence of specific derived requirements, any particular radar subsystem, or cascaded subsystems, be correctable to a coherence between channels be better than 99 percent over the entire band or any sub-band.

4.3 Imbalance Mitigation Techniques

Channel balancing is a system issue, and needs to be dealt with from a systems standpoint. Most importantly, mitigation of channel imbalances should not be an afterthought. It is a poor system design strategy to address channel balance with "Oh, we're not going to worry about it now, and just take care of it in software, later." Achieving good channel balance will generally require a collaboration of several techniques. We next examine a hierarchy of several techniques.

1. Prevent Imbalance

We note that any imbalance that can be prevented doesn't have to be dealt with later. Consequently, we begin with proper component selection, and circuit design/construction. This may necessitate, or at least be aided by, using matched components during module construction.[15] Once a circuit is constructed, the next technique is calibration with respect to parameters that can be adjusted, either with hardware or software. This might include gain, phase, and/or group delay. Such calibration might suffice with global values, or may be frequency dependent.

If the characteristic of imbalance holds over temperature, age, etc., then a single calibration will be adequate. However, if imbalance is not static, then repeated or iterative calibration might be required. In some cases, calibration needs to be adaptive to each and every data set. Examples of such adaptive algorithms include Adaptive Beamforming techniques and Space-Time Adaptive Processing (STAP) for GMTI systems. However, these is not always possible for achieving sufficient balance, and should not substitute entirely for more hardware-centric techniques.

2. Move Imbalance Energy

If imbalance cannot be eliminated, sometimes the imbalance energy can be separated from the balanced energy, and then filtered. For example, this can be achieved with channel commutation.[16,17]

3. Smear Imbalance Energy

Some aspects of imbalance might allow for imbalance energy to be smeared, or otherwise have its coherence suppressed to below the noise level in the data. One example of this might be random phase coding for each channel to decorrelate or decohere spurious signals or cross-talk.

4. Mitigate Effects

Our last line of defense is to mitigate the effects of channel imbalance in a downstream radar product, often after detection operations. Such techniques are often generically referred to as False Alarm Mitigation (FAM) techniques. Specifically, these might include channel averaging, or detection voting schemes.[18,19]

5 SUMMARY & CONCLUSIONS

In practice, frequently of greater concern than sibling receiver channels being ideal, is that channels exhibit like characteristics. Channel balance is of great concern to a number of radar applications, where we wish to remove the radar hardware/software influence on the relative measurements of various data streams. An excellent measure of channel balance between two channels is a measure of the ratio of their transfer functions. The ratio measure is related to (but not identical to) a coherence measure between the channels. In addition to gain, phase, and delay, channels should ideally exhibit the same noise floor and any nonlinear characteristics, as well as exhibit minimal crosstalk.

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