EVLA Memo 80

A Gain Slope Correction Scheme for the EVLA Receiver System

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Abstract : The EVLA Project Book specifies that the entire RF/IF signal chain must have less than 5 dB of gain slope and ripple across the 2 GHz input bandwidth provided to the Sampler. The Front-Ends are allowed slightly less than 2 dB of this gain flatness specification while the LSC and UX Frequency Converters and the final Downconverter are each allotted a similar amount. This is a very challenging requirement. The gain flatness of the first K and Q-Band receivers to be installed on the EVLA Test Antenna have been analyzed. It is evident that these wideband receivers will be unable to meet the 2 dB gain slope allotment, let alone the entire 5 dB spec at the band edges, without gain slope correction. A fixed gain slope equalizer will not be adequate since the receiver gain slope changes in both magnitude and direction depending on which part of the RF band has been selected. Also, the gain response will vary depending on which EVLA frequency band is being used and each of the 27 individual receivers in a particular band on the Array will be different. Thus a dynamic equalizer will be required. One possible solution to the problem is presented, based on a custom programmable equalizer which provides ±15 dB gain slope corrections in fixed 2 dB increments. The preferred location for this new equalizer, consisting of surface mount components on a microstrip circuit board, would be inside the T304 Downconverter module. The autocorrelation spectrum provided by the WIDAR Back-End would be used to determine the required gain slope correction.

Introduction:

Recently concern has been expressed that the EVLA baseline design for the RF/IF system may have unacceptable gain variation across the wide bandwidths that the EVLA will require. The Project Book specifies the overall gain flatness of the FE/LO/IF system to be less than 5 dB over any 2 GHz bandwidth with the design goal of 3 dB. This frequency range is set by the bandwidth of the high-speed digitizers. It is estimated that a 5 dB gain slope on the signal delivered to the 3-bit EVLA Samplers will lead to about a 3% loss in sensitivity. The essential problem is that as the power level of the signal drops, the quantization noise contributed by the digitizer starts to rise (gone are the good old days when astronomers were happy with a mere 65% efficient 1-bit sampler or the current VLA's 81% efficient 3-level sampler). The more bits the ADC has, the larger the gain "unflatness" we can tolerate. The 1 GHz, 8-bit Sampler which will be used mainly at the lower frequency bands to improve resistance to the effects of RFI, will have less of a problem with bandpass shape due to its higher dynamic range.

The EVLA's 5 dB gain variation over a 2 GHz bandwidth spec has been divided up equally between the 3 major RF/IF subsystems in the following manner:

- ⅓ from the Front-End
- ⅓ from the T302 (L/S/C-Band) or T303 (Ku/K/Ka/Q-Band) Frequency Converter
- ⅓ from the T304 Downconverter located directly in front of the Sampler

These are very tight tolerances to achieve over such broad bandwidths. The high-frequency receivers have ~10 GHz wide IF outputs while the subsequent IF converter chains have output bandwidths of 2 to 4 GHz. To put it into perspective, the gain slope achieved in the current VLA system is about 3 dB across its 50 MHz Sampler bandwidth. If this slope were extrapolated to reach the 2 GHz bandwidth used by the EVLA Sampler, it would be equivalent to a gain roll off of 120 dB. In other words, the VLA band shape requirements are relatively trivial compared to the EVLA's.

Since the EVLA Test Rack is still under development, we have no way yet to do a full-up test to see how the final 2-4 GHz Sampler bandpass is influenced by the various Front-Ends, the T302 & T303 Frequency Converters, the T304 Downconverters and their interconnecting cables. We can, however, look at the system which will likely be the biggest offender in the overall gain variation equation - the cryogenic receiver. Typically it is the active components in the signal path, such as the low-noise amplifier, the post-amp and the block conversion mixer, that define the shape of the RF passband. Often LNA designs have been optimized to achieve the lowest possible noise figure rather than provide perfectly flat gain across the entire bandpass. This may result in the gain response rolling off relatively steeply at one, or perhaps both, of the band edges. This is particularly true when the LNA's are used at frequencies beyond their optimum design bandwidth.

Another source of gain variation comes from the ripples arising from impedance mismatches in the RF or IF signal path. All of the receivers are designed with isolators in the most appropriate locations. That being said, it is almost impossible to eliminate the effects of all of the mismatch induced standing waves. Mixers can also be a major source of gain dips. Suck-outs in the frequency response can come from mismatches of the RF signal going into the mixer or the IF signal coupled out of the mixer. Mismatches on the LO port can also affect the drive level reaching the mixer which can increase the conversion loss. In the worst case scenario, all three can happen at the same frequency. Fortunately, the magnitude of these types of gain ripple usually cause less of a problem than the gain slope from the amplifiers. On the other hand, they cannot necessarily be corrected with gain equalizing circuits. The EVLA spec for passband ripple calls for a maximum of 0.2 dB for ripple with a period less than 2 MHz.

Analysis of a Typical K & Q-Band Receiver Gain Shape

Figure 1 shows a plot of the T_{Rx} and gain performance of the first EVLA-style K-Band receiver (S/N 27) as measured in the lab by the "Son of SOIDA" Test Rack. Both the LCP and RCP curves are presented. The T_{Rx} axis is on the left of each graph while the system gain axis is on the right. The data points were recorded in 100 MHz frequency steps from 18 through 26 GHz. The data was measured in two ways. In the Swept LO1 mode, shown as red traces, the first LO is moved in 100 MHz steps from 29 to 37 GHz while the 2nd LO is held fixed at 11 GHz. This effectively allows us to mix a 40 MHz bandwidth about the desired RF frequency down to a constant 11 GHz IF and then convert it to baseband where it is detected by a power meter. In the Block Converter mode, shown as blue, the first LO is fixed at 34 GHz, resulting in the entire 18-26 GHz RF block being converted down to an 8-16 GHz IF (actually 16-8 GHz since there is a spectral inversion). The 2nd LO is then stepped along in 100 MHz intervals to mix the 40 MHz portions of the 8-16 GHz IF down to baseband. Each method used "hot" (ie: ambient) and "cold" (ie: LN2) loads to determine the Y-factor across the full receiver passband.

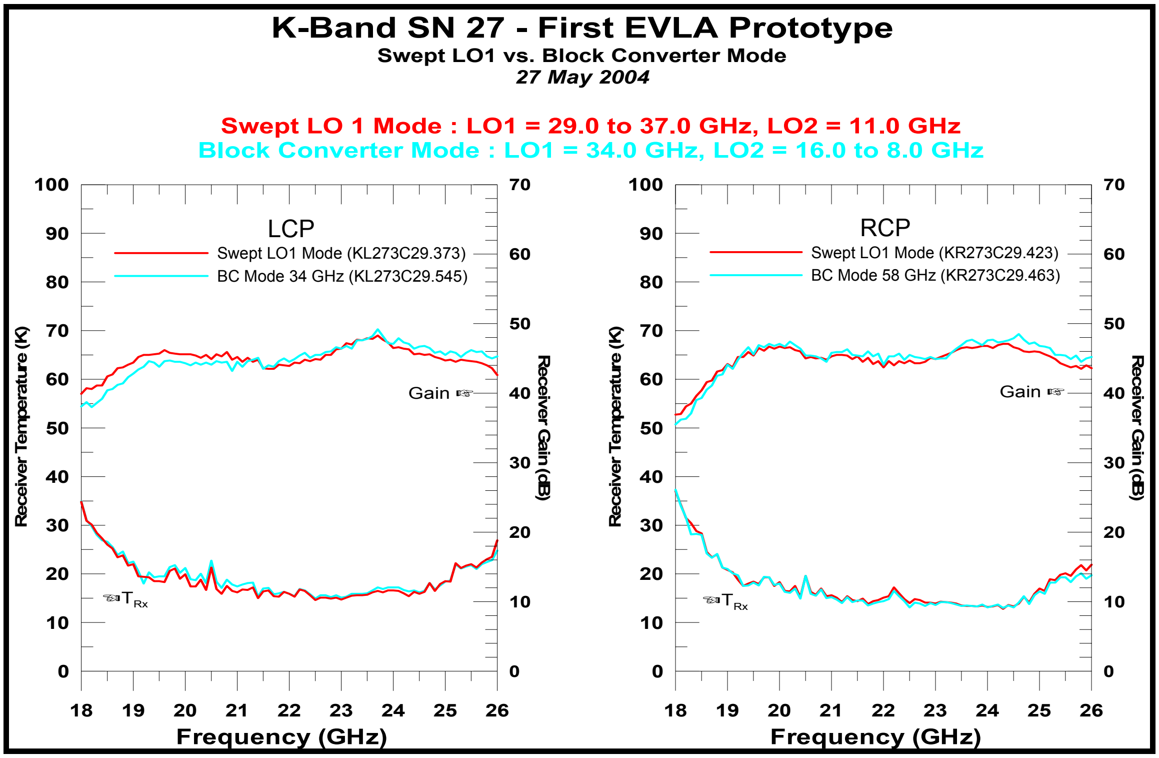

Figure 1 : Gain and T_{Rx} for the first EVLA K-Band Receiver

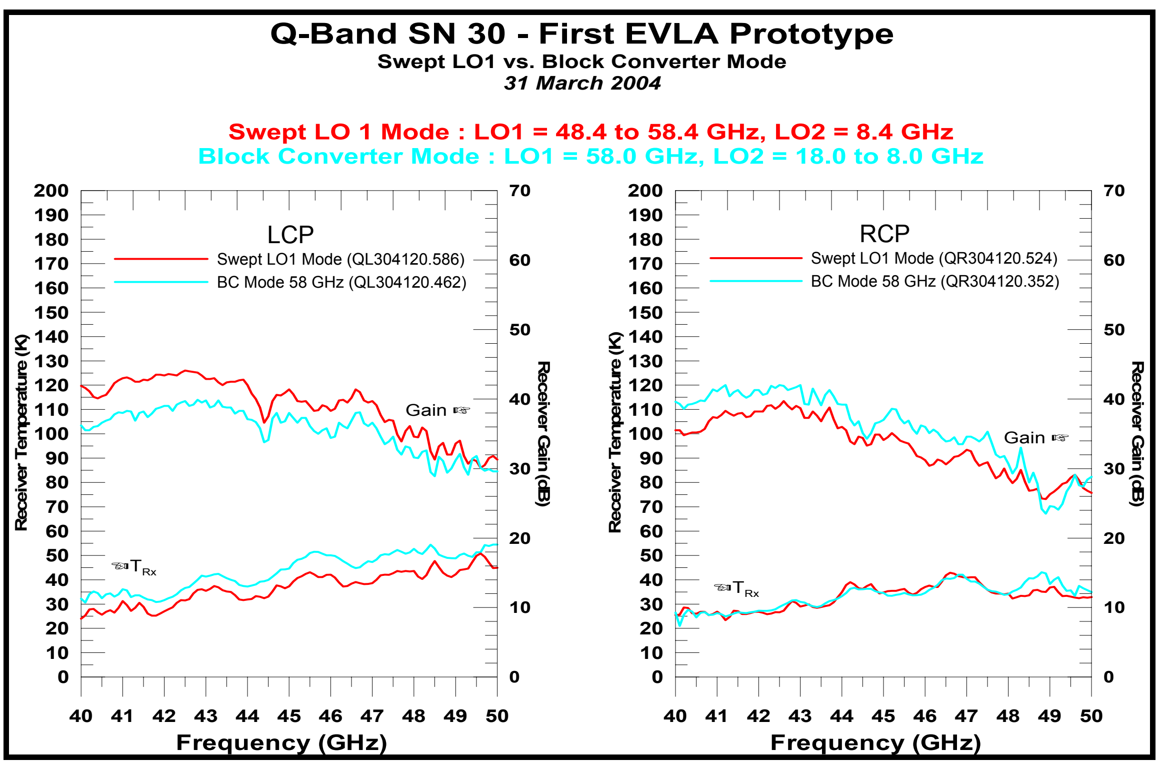

Figure 2 : Gain and T_{Rx} for the first EVLA Q-Band Receiver

Figure 2 shows the performance of the first EVLA-style Q-Band receiver (S/N 30). For testing this Front-End, the Swept LO1 method used a fixed IF frequency of 8.4 GHz while the 1st LO is swept across 48.4-58.4 GHz in 100 MHz steps. In the Block Converter mode, the 1st LO was set to 58 GHz, which downconverted the entire 40-50 GHz RF band to an IF of 18-8 GHz (again, there is a spectral inversion from using a high side LO), where it is then mixed down to baseband by stepping the 2nd LO across 18-8 GHz in 100 MHz steps.

While the Swept LO1 and Block Converter modes essentially show the same thing, the two techniques provide different insights into receiver performance. The Swept LO1 scheme eliminates the effect of the gain response from the Base-Band Converter (BBC) since it uses a fixed IF frequency (11 GHz in the case of K-Band, 8.4 GHz for Q-Band). However, it may result in noticeable gain effects from the 1st LO as it is swept across its full frequency range. Mismatches in the LO chain, especially in the frequency multiplier (ie: K-Band uses a doubler while Q-Band uses a tripler) or in the mixer, can cause suck-outs or LO induced gain ripple.

The Block Conversion mode, on the other hand, will impose the frequency response of the BBC onto the receiver's gain curve. Since the 1st LO is stationary, there will be no frequency dependence in the gain from LO power variations, although the overall conversion loss of the mixer may increase if the fixed LO happens to fall on a frequency where there is a multiplier or mixer suck-out. This is why the T_{Rx} curve seen on the LCP side of the Q-Band receiver in the Block Converter mode (Figure 2) appears degraded compared to the Swept LO1 mode. The Block Converter mode is very similar to the configuration that will be used on an EVLA telescope. However, in a full-up K or Q-Band receiver system we would also see the gain slope across 8-18 GHz IF caused by the cables connecting the Front-Ends to the T303 UX Converter, as well as the gain response of the UX Converter itself. The cables connecting its output to the T304 Downconverter will introduce an additional 8-12 GHz gain slope, followed by the gain response of the Downconverter. The Sampler, of course, will contribute its own frequency response to the mix.

The gain response exhibited by both polarizations in Figure 1 and 2 show undulating gain curves and are far from being flat. The low-end of the K-Band receiver has a very steep roll off in gain and has a pronounced double hump. The gain on the Q-Band receiver slowly rolls off with increasing frequency. These effects are primarily due to the LNA's and other waveguide components in the RF path. Its obvious that the front-end systems alone could easily exceed the entire 5 dB flatness specification even ignoring the extra gain slopes that will be added by the IF cables and Converters. However, it must be remembered that the amplifiers employed in our K & Q-Band receivers are versions of the LNA's developed for the MAP Project and while they have been modified with Cryo-3 devices to achieve better noise performance, their optimized bandwidths are still based on the original design for NASA (ie: the 20-25 GHz and 35-46 GHz MAP bands). Consequently, we are asking the LNA's to work well outside their design spec (and they do so delightfully well).

To show what the gain slopes look like in greater detail, the 8 GHz bandwidth of the K-Band receiver has been carved up in four 2 GHz wide chunks and plotted in Figure 3. The gain curves have been normalized by subtracting the average value (the T304 Downconverter has programmable attenuators which are used to adjust the average power level into the Sampler module to be -29 dBm, which is essentially the same process being simulated here). As can be seen, the slopes across most of the 2 GHz wide chunks wander well outside the ±1 dB box used to show a liberal interpretation of the ⅓ share of the 5 dB EVLA flatness spec (ie: 1.7 dB).

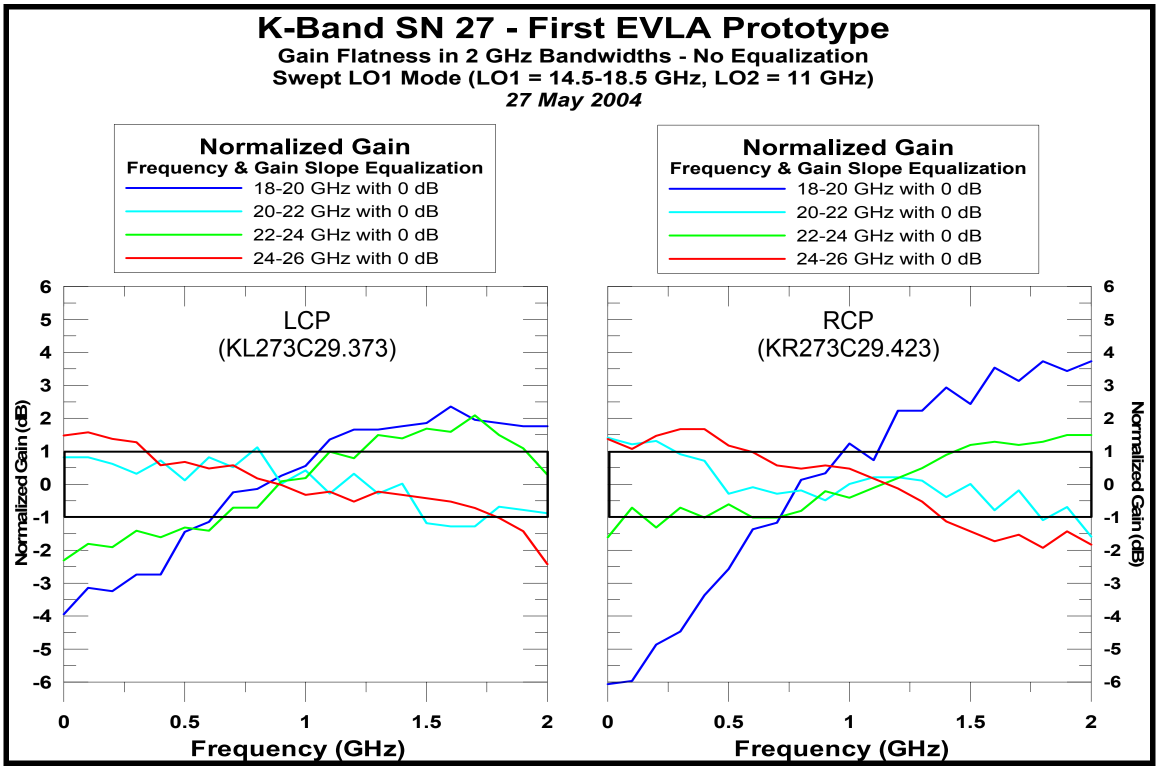

Figure 3 : Normalized Gain Across 2 GHz Bandwidths at K-Band

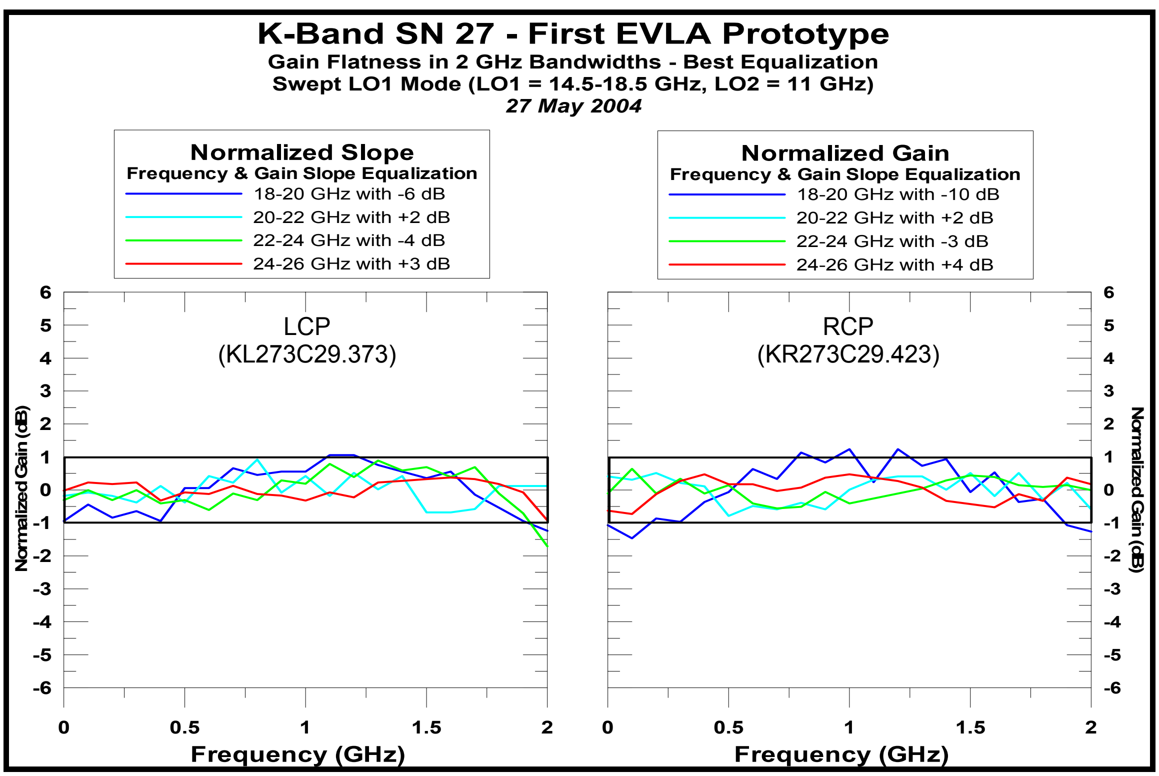

Figure 4 : Best-Case Gain Equalization at K-Band

It is thus obvious that we will need a gain equalizer of some sort to compensate for these excessive slopes. Unfortunately a fixed equalizer won't be of much help since the gain slopes change in both magnitude and direction depending on which part of the RF band is being looked at. And no two receivers will be identical. The gain slopes exhibited by each of the 27 K-Band receivers on the EVLA will likely be different. And obviously each receiver band with 2 GHz or more of bandwidth (ie: S, C, X, Ku, K, Ka & Q-Band) will have its own distinct passband shape.

The ideal equalizer would be a filter which has a single pole that can be moved around in the frequency domain. A continuously variable equalizer would allow us to compensate for both negative and positive gain slopes. A SAW filter might do this, but the 2 GHz bandwidth would complicate such a design. A much more brute force attempt to remedy the problem was made by assuming we had a programmable equalizer with gain slopes that could be selected in fixed 1 dB increments across the 2 GHz bandwidth with both positive and negative slope compensations available. By choosing an equalizer setting which yields the lowest variance for the gain curve, the linear component of the gain slope can be removed. Figure 4 shows the resulting improvement in the K-Band receiver. As you can see, we now come very close to fitting within a 2 dB window.

A similar process was done for the Q-Band receiver. Figure 5 shows the 10 GHz wide gain curve for Q#30 broken up into five 2 GHz chunks, normalized and plotted on top of each other. Although there is much more gain ripple in the Q-Band Front-End than in the K-Band receiver, most traces do exhibit a noticeable positive or negative gain slope. If we had an idealized fixed equalizer with 1 dB slope increments which could be used to compensate the linear gain shape, it would yield the curves shown in Figure 6. While not all of the traces fall within the 2 dB window, the flatness is generally much improved.

Those traces that exceed the desired spec are usually from gain dips or suck-outs. Just why the Q-Band receiver has much more character to its bandpass is not known at the moment. These receivers have always shown larger ripple and gain suck-outs although it does not usually affect the receiver temperature. This indicates the mismatch (or mismatches) must arise after the LNA since a 2 dB gain dip in front of the first gain stage could nearly double the T_{Rx}. Consequently the mismatches must be arising in the waveguide run between the LNA and post-amp or between the post-amp and the mixer or between the mixer and the receiver output. However, there are isolators already located in the signal path in these locations so it may be difficult to effect any significant improvement in the gain flatness over what appears in Figure 6. However, the Q-Band ripple will be explored further on the next EVLA receiver.

A Simplified Programmable Gain Equalizer Scheme

While a fixed equalizer with 1 dB slope increments would certainly be adequate for correcting the gain slopes expected to be encountered in the EVLA, such a programmable equalizer might not be all that easy to build because of its large number of required settings. For example, an equalizer with selectable slopes between ±10 dB would require an RF switch with 21 positions. The simplest design which we might consider would be one that provides slopes ranging from -9 to +9 dB in 3 dB increments. Figure 7 shows a block diagram of this bare bones equalizer showing its 7 programmable gain settings.

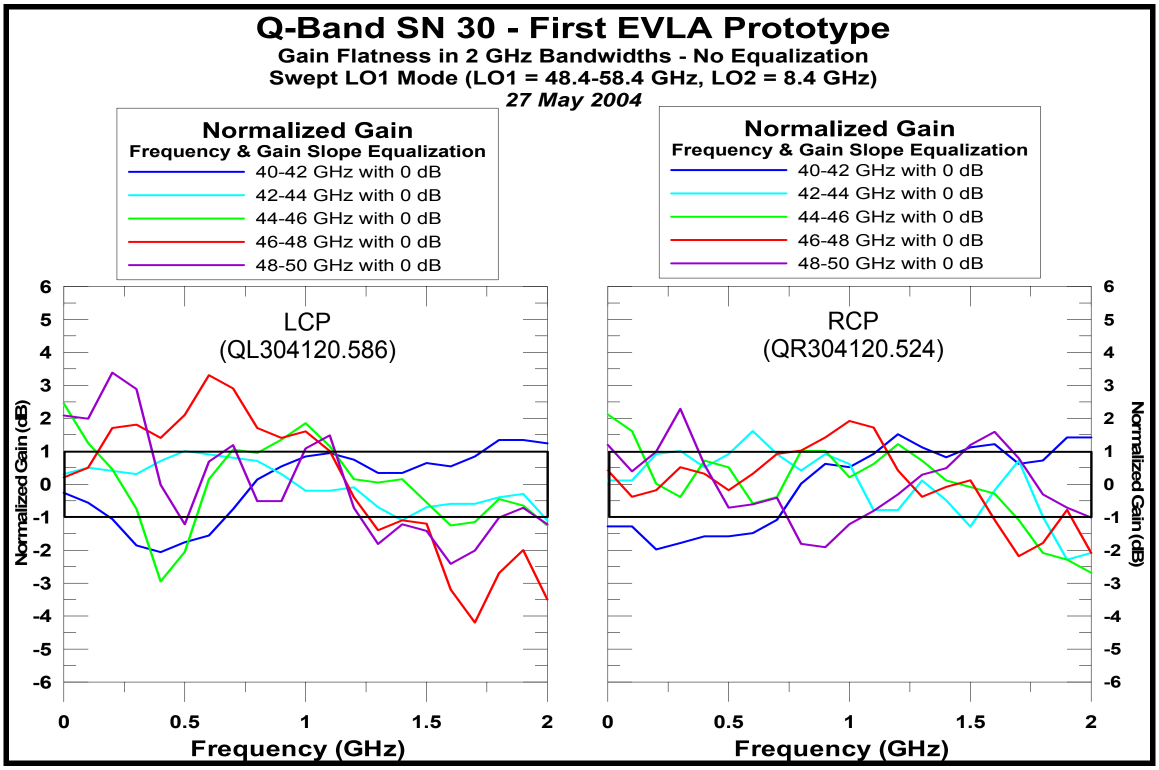

Figure 5 : Normalized Gain Across 2 GHz Bandwidths at Q-Band

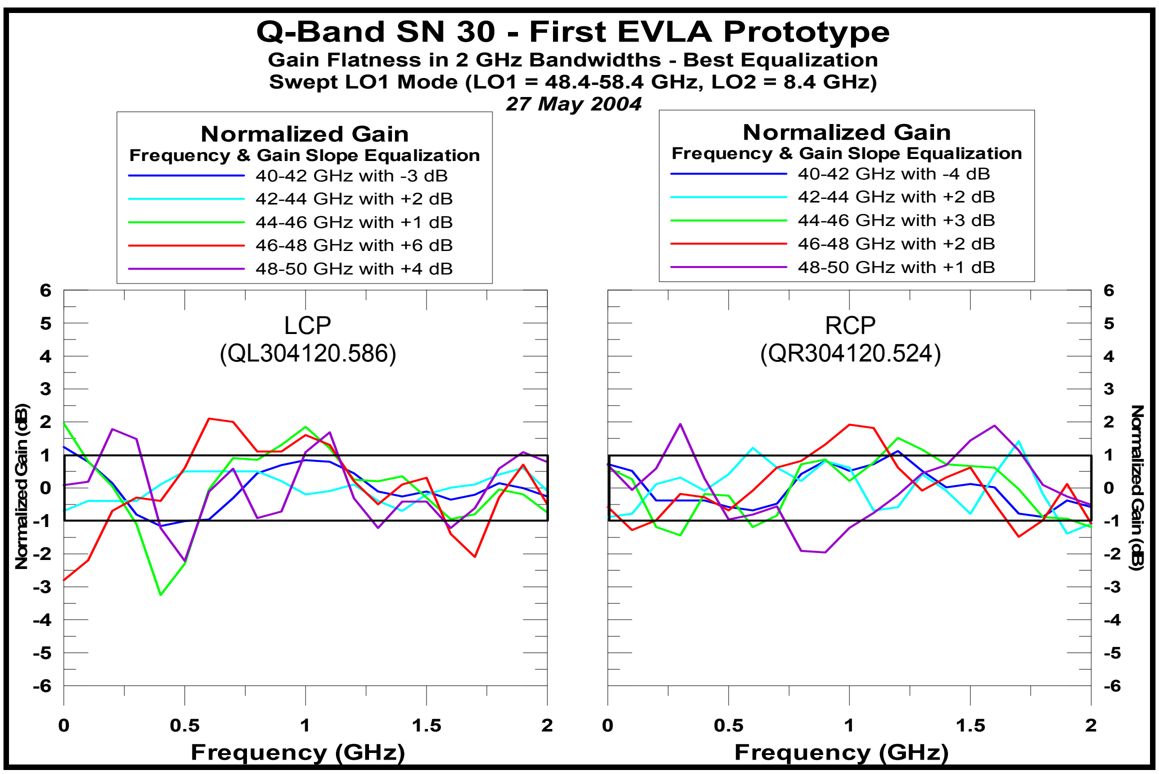

Figure 6 : Best-Case Gain Equalization at Q-Band

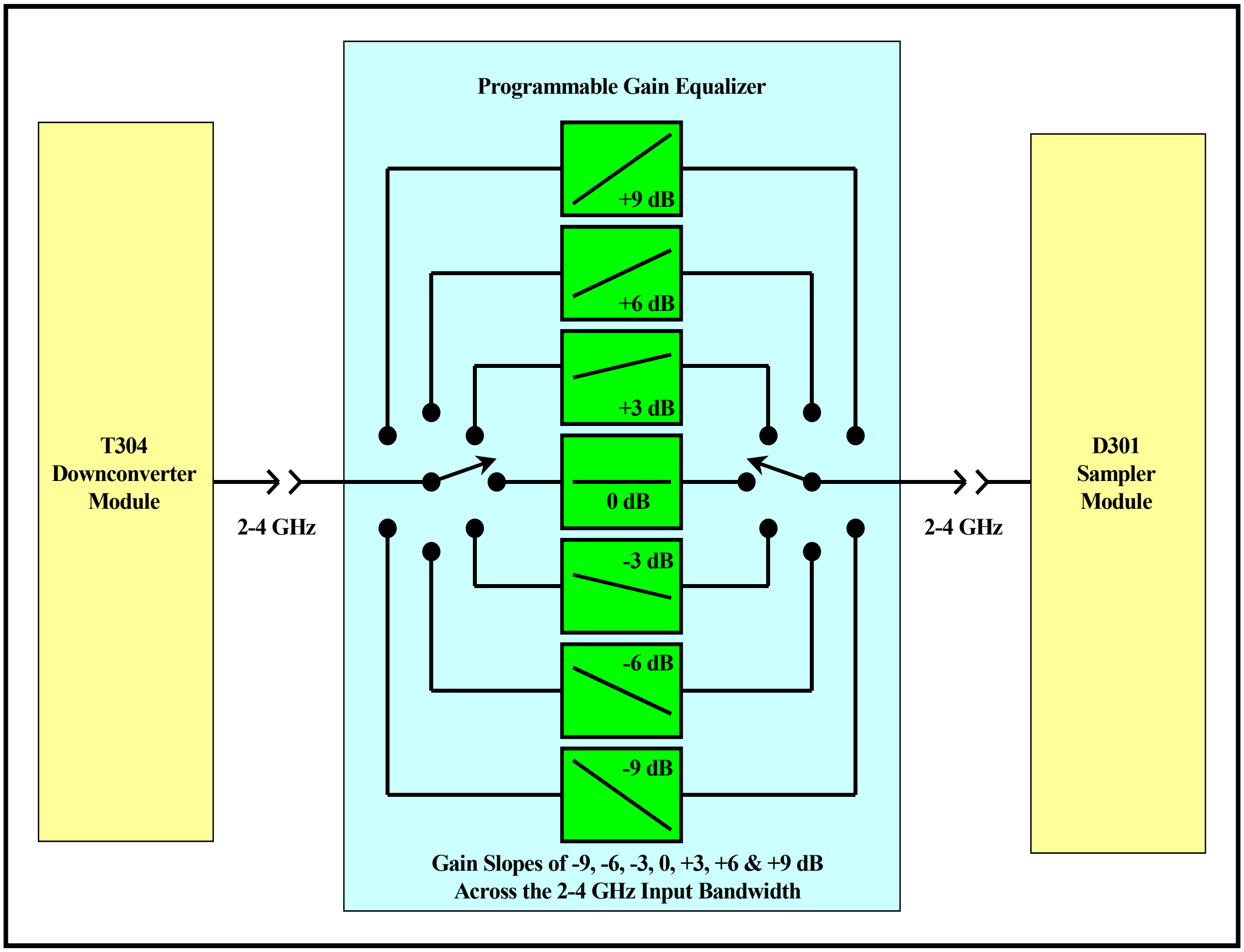

Figure 7 : Simple Programmable 2-4 GHz Gain Slope Equalizer

The unit would be located between the T304 Downconverter and the D301 Sampler, or perhaps within the Downconverter itself, and would be designed for an operating frequency of 2-4 GHz. A pair of multi-position switches are used to make the switch ladder which allows the desired gain correction setting to be selected. While a programmable equalizer could be inserted further upstream in the signal path, either in the 8-18 GHz input to the UX Converter or the 8-12 GHz input to the Downconverter, placing it in front of the Sampler is preferred because of the lower frequency and narrower bandwidth.

To evaluate how well a programmable equalizer with 3 dB slope increments would work compared to the 1 dB step version, the normalized gain curves were revised to use the closest 3 dB step. Figure 8 shows the K#28 receiver. The coarser slope correction is only slightly worse than the finer resolution equalizer shown in Figure 4. Similarly, Figure 9 shows the Q#30 receiver using 3 dB steps. Contrasted with Figure 6, there is little degradation seen with the coarser step size. Its flatness is still dominated by the receiver's gain ripple and suck-outs rather than the gain slope.

While this is only a limited sample, it suggests that even this relatively simple programmable equalizer could go along way to compensate not only the gain slope arising in the Front-Ends but the entire signal chain, making the 3 dB flatness design goal over any 2 GHz bandwidth within reach.

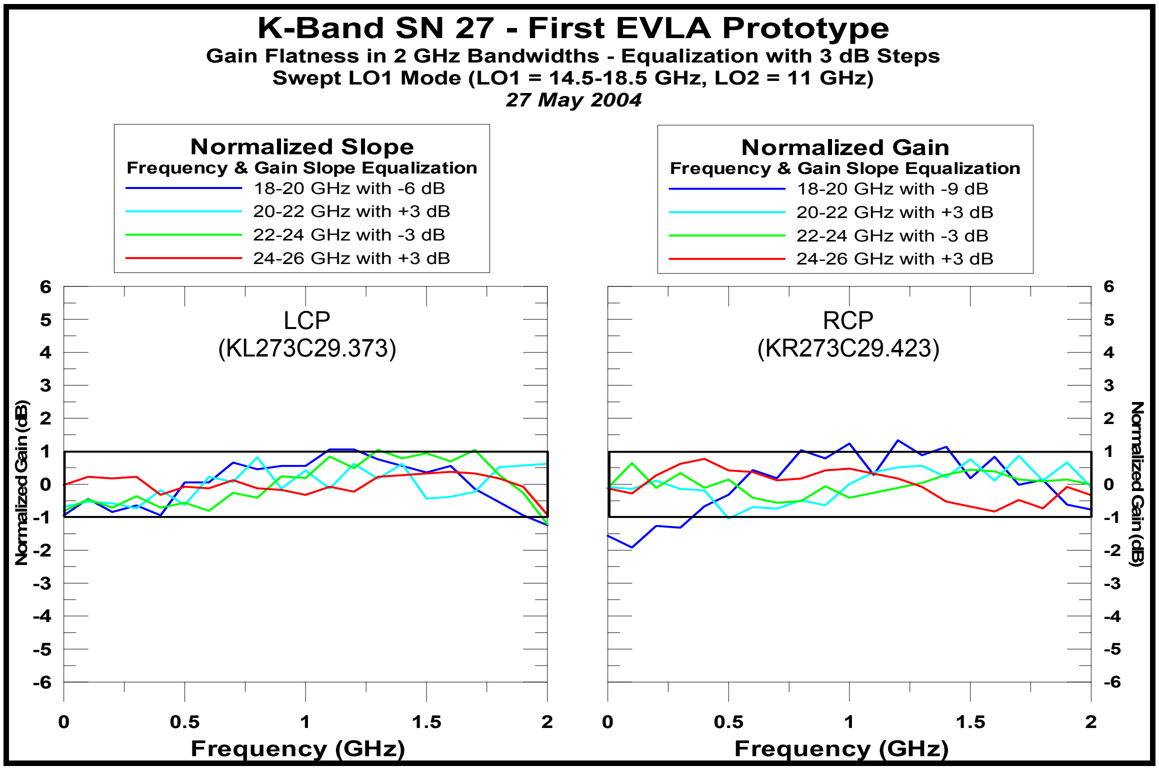

Figure 8 : Gain Correction at K-Band using an Equalizer with 3 dB Steps

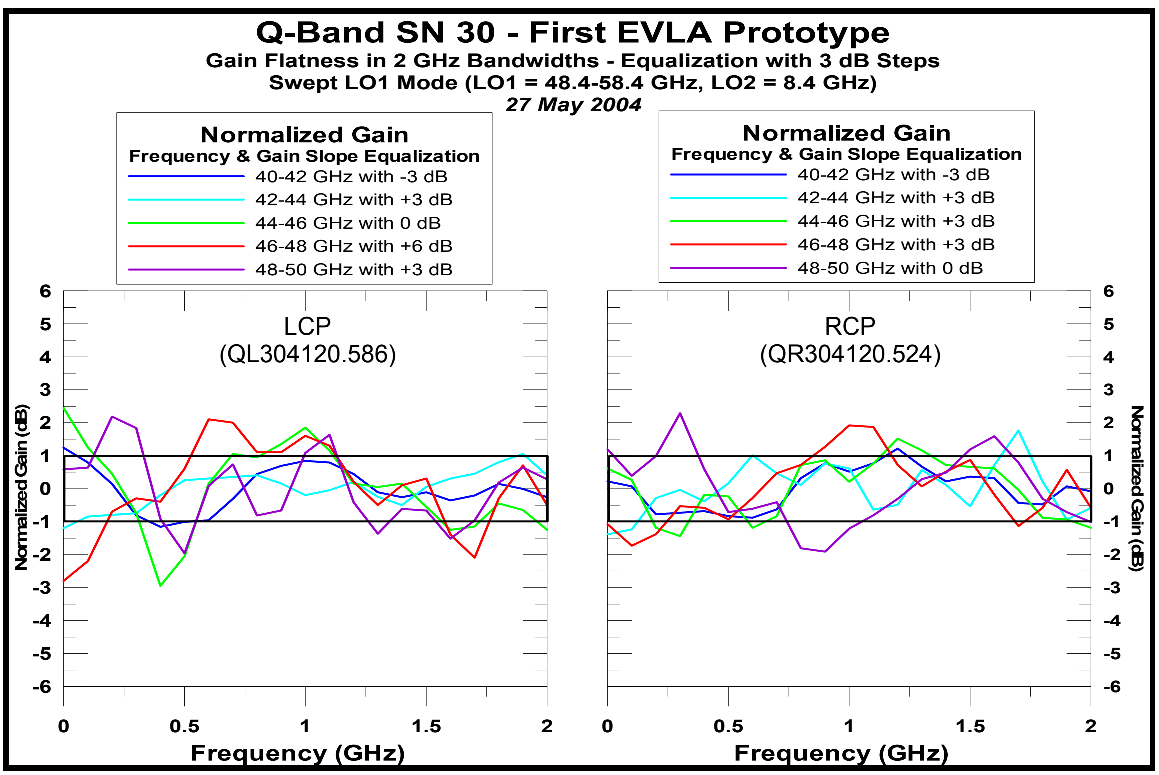

Figure 9 : Gain Correction at Q-Band using an Equalizer with 3 dB Steps

Implementation Details

While the programmable equalizer scheme has the potential to flatten the frequency response across the 2-4 GHz bandwidth of the signal delivered to the Sampler, there is still the problem of knowing exactly what gain slope correction should be applied. One possible solution is to use the auto-correlation feature provided by the WIDAR Correlator. As well as processing all of the cross-correlations for each interferometer baseline, WIDAR will also be able to measure the auto-correlation of the signal associated with each element in the array, the Fourier Transform of which gives us the power spectrum of the 2 GHz wide band digitized by the Sampler. Measuring it in this fashion will give us the total gain slope including the contributions from the entire signal path (ie: Front-End, Converters and cables, as well as the Sampler's response). It should be possible to determine the gain slope by fitting a straight line across the mid 80 to 90% of the band (ignoring the band edges). The desired equalizer setting would simply be the closest slope of the opposite sign (ie: if there was a -4 dB bandpass slope, we would chose the +3 dB equalizer setting).

If this scheme was adopted, it would obviously be an added task for the real-time EVLA system software. It would be complicated by the fact that the 2 GHz bandwidth is divided up into a number of sub-bands by the digital FIR filters in the Correlator. The 2 GHz wide power spectrum would have to be knitted together before the linear fit could be performed. An alternative scheme would be to use lookup tables to determine the required compensation slope. However, this could be difficult to implement operationally. The magnitude and sign of the required correction will vary with the receiver band in use, each receiver in that band and the particular frequency it is tuned to, as well as the idiosyncracies of each of the LXC/UX and Downconverters. Every time the Front-End or the Converter on an antenna was replaced, the lookup table would have to be modified. Measuring the gain slope directly with the Correlator would ensure the equalizer setting was always correct.

Since this scheme has the potential to correct for all of the gain slopes in the RF/IF/LO system (but not for gain ripple), it has a good chance of ensuring we will be within the 5 dB flatness window. However, inserting an equalizer into the signal will drop the overall power level into the Sampler significantly (by about ½ of the gain slope value plus the fixed insertion loss of the switch ladder). Consequently, the digital attenuator in the Downconverter will have to be readjusted to ensure the input level to the Sampler is close to its optimum -29 dBm level. Without having tried this closed loop method in practice before, we may find that we will have to iterate several times to get the flattest response, that is, measuring the slope, setting the equalizer, changing the attenuator, remeasuring the slope, resetting the equalizer, etc.

While inserting the programmable equalizer between the Downconverter and the Sampler would provide us with an adequate method of controlling the gain slopes, it may not be the best place in the signal path to locate it. Figure 10 shows a block diagram of the T304 Downconverter baseline design. When set to any of the steeper compensation slopes, the equalizer will introduce a significant insertion loss. Although the output digital attenuator likely has adequate range to bring the level back up, it is possible that, should the equalizer be located directly in front of the Sampler, the output amplifier in the Downconverter could be driven into compression. Accordingly, it would be advantageous to integrate the equalizer into the Downconverter and place it in front of the output digital attenuator as well as the output amplifier. This would have the added benefit of having the total power detector measuring the actual output power level after equalization takes place.

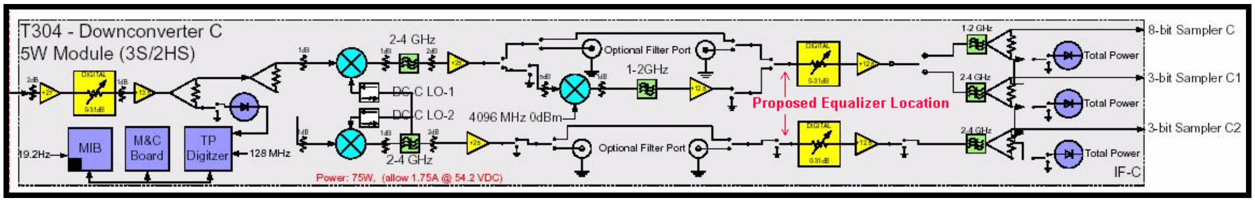

Figure 10 : Block Diagram of T304 Downconverter Baseline Design

The Downconverter is designed with an "Optional Filter Port" which would be the ideal place to test out the scheme before committing to a more permanent location. Since it would be a shame to commandeer the T304's only available auxiliary signal path, it would make sense for the final version of the Downconverter to have it placed between the Optional filter output switch and the digital attenuator (indicated by red arrows in Figure 10). As will be described in the next section, the space required for the equalizer circuitry, which would use surface mount resistor, capacitor and inductor devices, would more than likely fit on the integrated Downconverter board currently being designed.

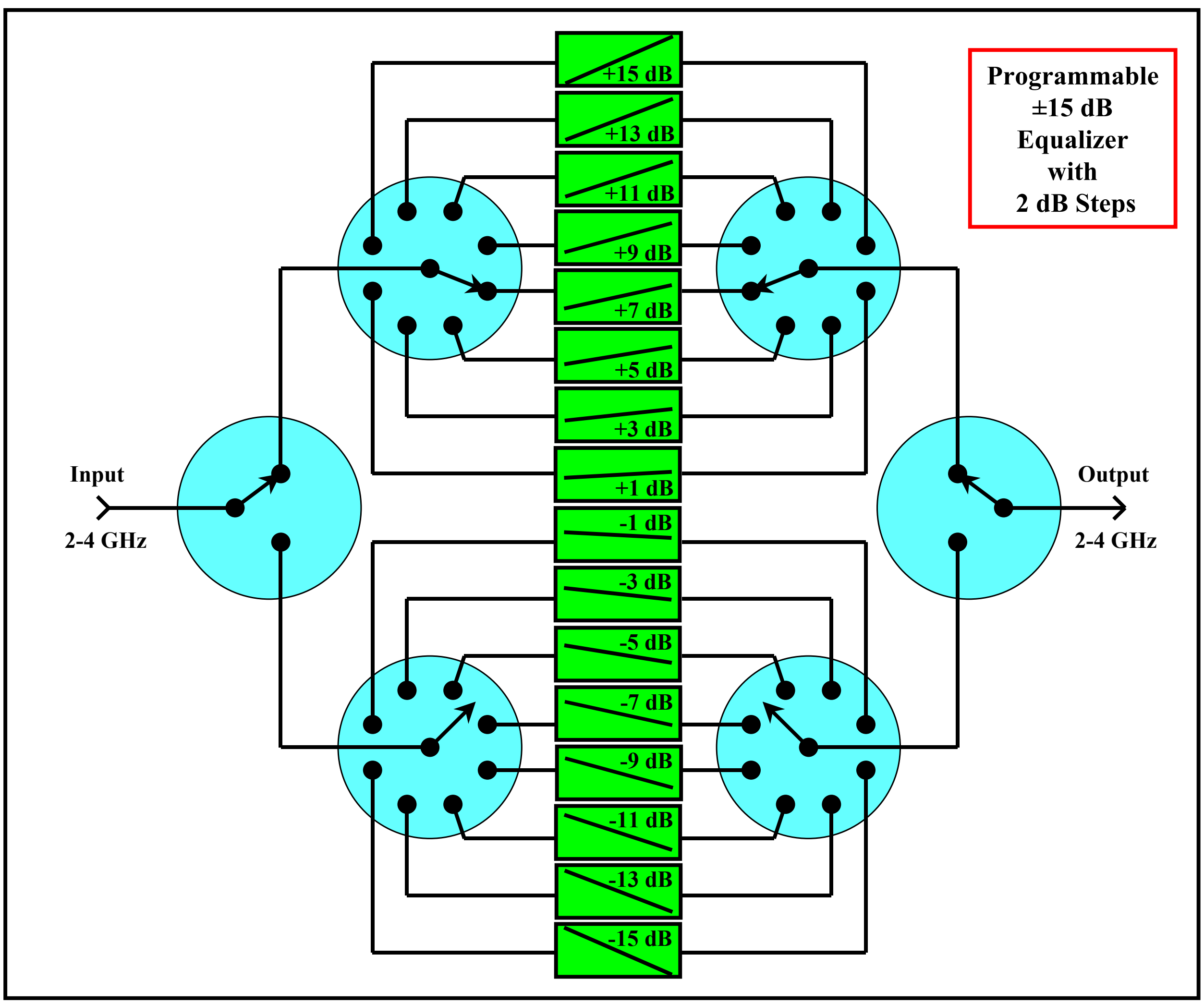

Figure 11 : Proposed Programmable ±15 dB Equalizer with 2 dB Step Size

For the two receivers looked at in the analysis, a programmable equalizer with fixed gain slopes between ±9 dB seemed adequate. However, as the effects of the frequency converters and cables were ignored, it is possible that the maximum compensation slope might need to be even steeper. For example, in the case of K#28, we would need the maximum negative slope setting (ie: -9 dB) to correct the receiver's passband at the low end of the band. Even if both the UX Converter and the Downconverter were within their 2 dB flatness allotment, Murphy's Law would inevitably conspire to ensure that both of their slopes are in the same direction as the receiver's sub-band, which would mean that we might need at least -13 dB of equalization.

The switch ladder configuration shown in Figure 7 could be extended to give us both steeper slopes and finer slope increments. A possible layout for the latter is shown in Figure 11. A sixteen rung switch ladder is used to implement a ±15 dB slope equalizer in 2 dB steps (ie: -15, -13,.., -1, +1,.., +13, +15 dB). Since we will likely be limited to single-pole 8-throw surface mount switches, the ladder will require a pair of 8-ways switches selected by a 2-way switch on both the input and output ports. The impact of the total insertion loss of the switch network, which could be as high as 8 dB, would have to be analyzed further to ensure the Downconverter has adequate gain. Also, with such steep equalizer slopes, it is possible that the equalizer circuits might need additional gain as well.

It would be desirable to come up with a basic programmable equalizer design that could be adopted by ALMA and the GBT as well as the EVLA. The suitability of the scheme proposed here should be evaluated for all 3 instruments.

Phase Considerations

According to the EVLA Project Scientist, the different states of the equalizer need not have identical phase characteristics. Ideally the phase in all states should be "delay-like" (ie: the phase is linearly proportional to frequency). While this type of phase characteristic would be convenient, it is not necessary since the Correlator can adjust for the equivalent delay as needed. Assuming good stability, these slopes could be determined and applied a-priori through periodic test measurements. These delay corrections would be transparent to the observer. The limiting phase slope, as specified in the EVLA Project Book, is 1 degree per MHz. This is a fairly lenient requirement. Anything smaller than that can be removed by the Correlator without any loss of closure.

Ripples and other non-linear phase or amplitude effects would be calibrated out by the observer through the standard procedure of observing a calibrator source. These phase restrictions may be more of an issue for ALMA since there are fewer strong cal sources available at sub-mm wavelengths to use. The most significant concern for the EVLA is that any phase structure versus frequency in the passband must be smaller than the frequency resolution of the Correlator or it could affect phase closure.

Finally, although the restrictions on the phase properties are fairly relaxed because of the EVLA's capability to calibrate them out, there is one absolute requirement - they must remain stable with time. These stability specifications, however, are no more stringent than those placed on any of the other components (ie: amplifiers, filters, attenuators, etc.) that we are planning to use.

Equalizer Design

A number of equalizer design approaches have been discussed at the NTC. Several engineers have suggested that at lower frequencies one possible solution would be to use an op-amp with the appropriate feedback impedance to give a sloped response. A variable feedback network might be designed to allow either continuous or discrete tuning of the gain slope. Op-amps in the traditional sense are clearly not available in the 2-4 GHz frequency range, but it may be possible to develop a differential microwave amplifier with feedback to achieve similar results. A significant benefit of this approach is that it could in principle have 0 dB mid-band loss (or even gain) regardless of the magnitude of slope correction that needs to be applied. While this would certainly be an elegant solution, the implementation could be very difficult. It is not clear what kind of feedback network could be designed that would 1) be compatible with commercially available differential amplifiers, 2) provide a roughly flat gain slope over an octave bandwidth with ±15 dB tunable range, 3) maintain reasonable impedance match on both input and output, and 4) ensure stability of the amplifier in all potential feedback states. It seems apparent that this novel and complex design would require a major R&D effort. If the resources were made available to develop such a concept, it might yield an attractive alternative for future projects.

A more straightforward and less risky approach for achieving a sloped frequency response is with passive, low-order filters. As stated earlier, a single-pole filter would be sufficient to provide both positive and negative gain slopes of almost any magnitude. In principle, this could be implemented in either a continuously-tunable fashion, or as a switch-ladder network of fixed filter designs. However, one must consider how the large slopes required will aggravate mismatch effects in the system. The preceding analysis of receiver data has shown that gain slopes of *at least* 9 dB will be needed from this equalizer, and probably as large as 15 dB. A classic lossless filter with 9 dB insertion loss will typically have a VSWR of almost 30:1, and a 15 dB slope filter would have a VSWR over 120:1! The equalizer would thus be extremely sensitive to impedance mismatch on both the input and output. Even a moderate mismatch from components adjacent to the equalizer could deflect its insertion loss by several dB.

Isolators could be used to ease the mismatch problem, but in addition to being bulky this would come at great cost to the project. We would need at least 2 isolators per band, or 16 per antenna. At approximately $200 per isolator, that would mean spending about $100K over the whole project, just for the isolators alone!

A better solution is to use non-reflective filters that dump the excess energy at the high-loss end of the band into resistors, rather than reflecting it back to the source like an ordinary filter would. An example of such a structure is shown in Figure 12. Resistors R1 and R3 are shorted out by the series LC resonators at mid-band for the filter, while resistor R2 is open-circuited by the parallel LC resonator. Off-resonance, these three resistors form a tee-attenuator. An example of the frequency response for such a structure that can achieve large gain slopes is shown in Figure 13. As can be seen, with the filter centered at 4 GHz, a significant positive slope is achieved across the 2-4 GHz range, while a return loss of better than 25 dB is maintained at all frequencies. Note also that the minimum loss is 0 dB at 4 GHz, illustrating that this "lossy" structure will not introduce any greater insertion loss to the system than any other passive network that has a sloping frequency response.

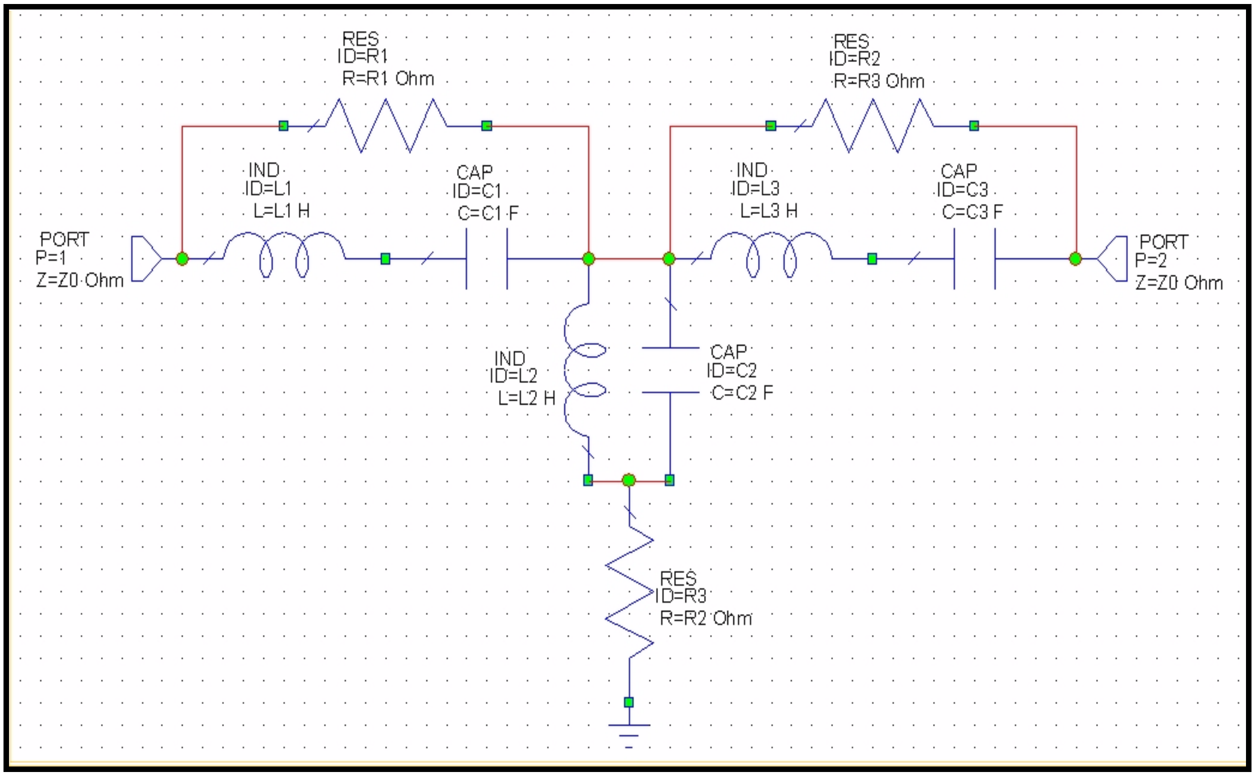

Figure 12 : General topology of the non-reflective filters.

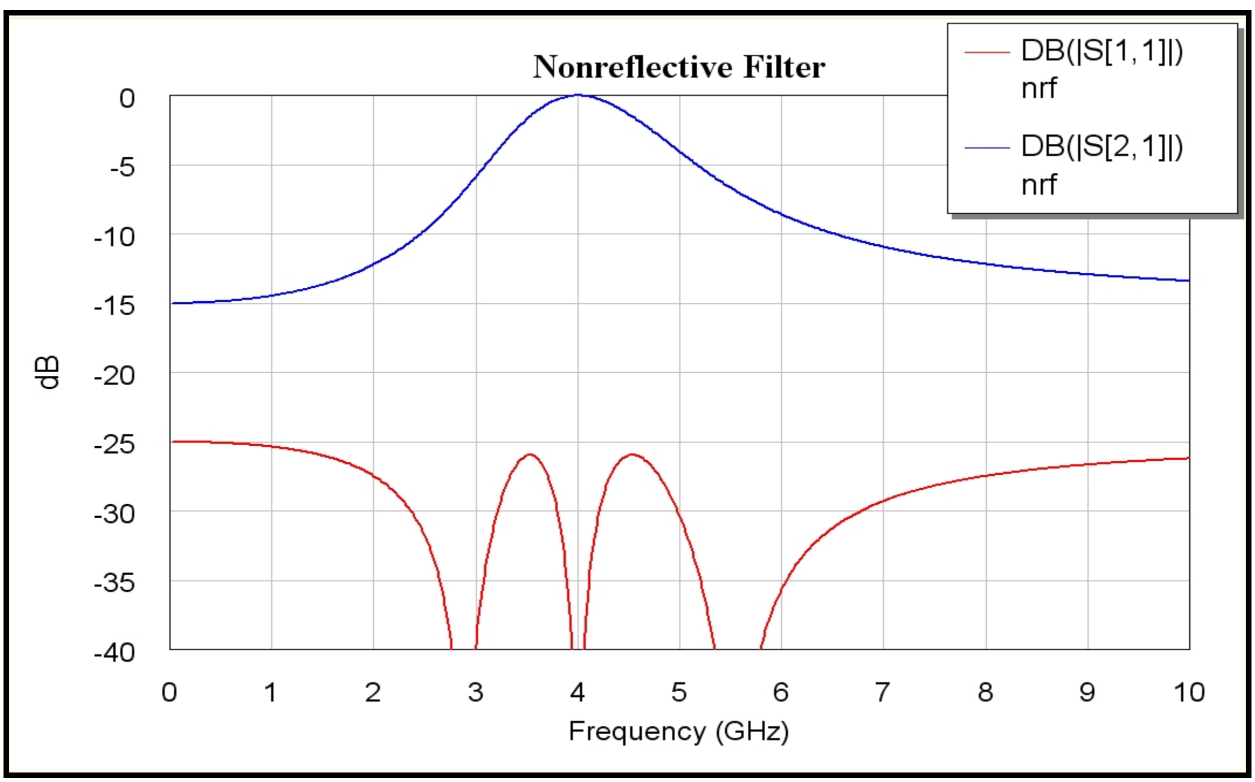

Figure 13 : Typical broadband frequency response of the non-reflective filters.

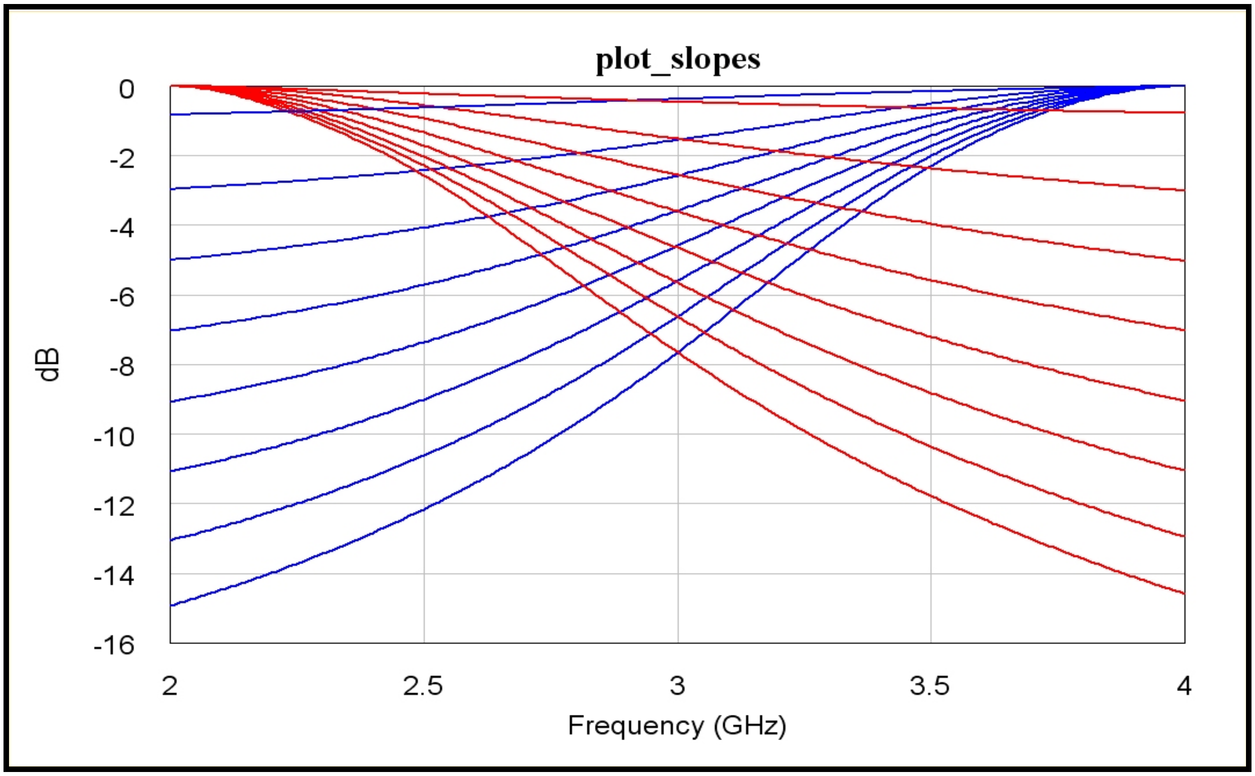

Figure 14 : Insertion losses of the proposed filters with slopes ranging from ±15 dB in 2 dB increments. Does not include the loss of the switches or embedding transmission lines.

It was decided for simplicity that the non-reflective filter described above would be implemented as fixed elements in a switch-ladder network. This would also tend to have better phase stability than a continuously-tunable structure dependent on analog controlled bias points. As was described earlier, a programmable slope equalizer with ±9 dB range in 3 dB steps would be considered a bare minimum for the EVLA, which could be achieved with two SP8T switches, one on the input and one on the output (SP7T switches are not commonly available).

However, the system may need as much as ±15 dB range, in which case it will be necessary to have at least 11 rungs. An extensive search on commercially available switches was conducted to determine what size and configuration the ladder network should be. No switches are available at this frequency in a compact package with more than 8 positions, so one must consider at least a two-level cascade of switches. Reasonable RF performance can be achieved with either an SP4T-SP4T or SP2T-SP8T configuration. The latter option was chosen because it uses fewer components.

Since the chosen ladder network gives us 16 rungs to work with, we can design the fixed filters to cover -15 dB to +15 dB in 2 dB steps, as shown in Figure 14. In these examples, the lumped element values were optimized to give the desired total slope with the straightest line possible while maintaining return loss better than 20 dB (see Figure 15). The phase responses are shown in Figure 16. Although they do not exhibit "delay-like" characteristics (ie: the phase does not vary linearly with frequency), the phase responses are smooth with gentle gradients and would likely be monotonic when combined with the other phases in the rest of the system, and thus should be easy to calibrate out using standard VLA calibration techniques.

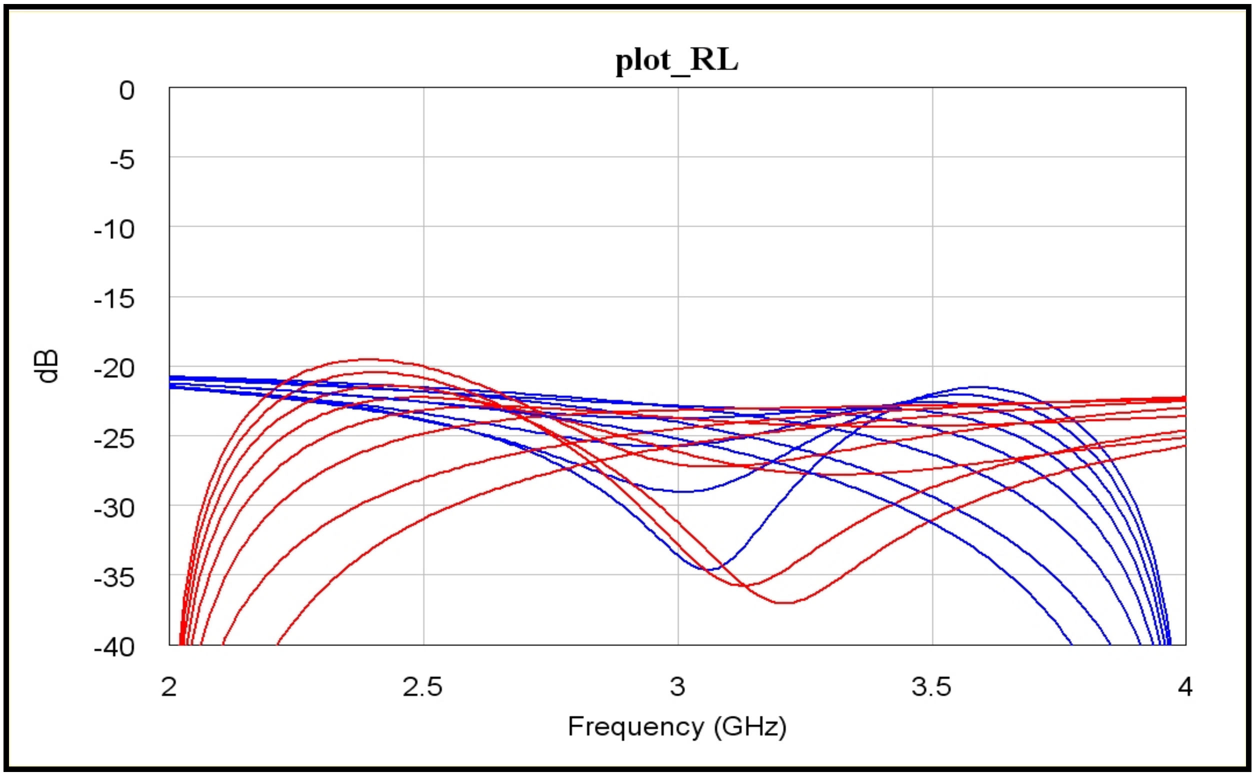

Figure 15 : Return loss of the proposed set of filters. Red curves are for negative-slope filters, and blue curves are for positive-slope filters.

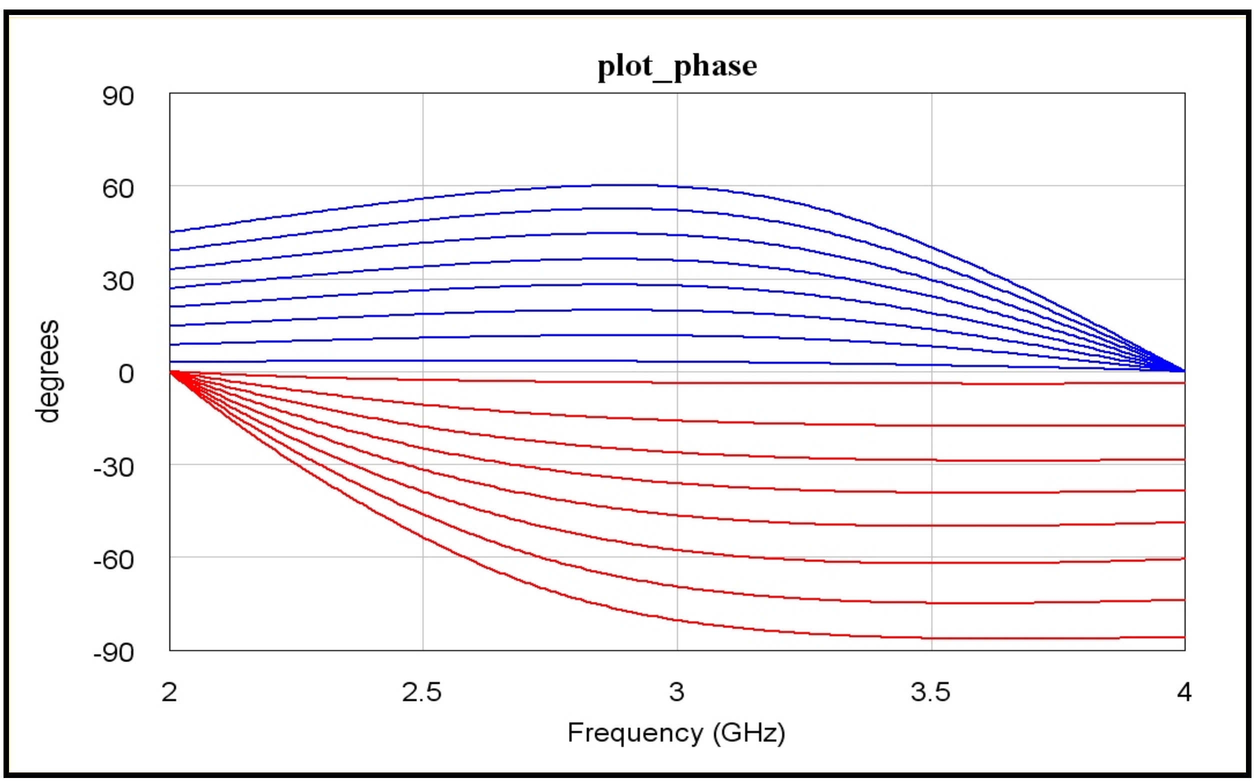

Figure 16 : Phase delay of the proposed set of filters. Red curves are for negative-slope filters, and blue curves are for positive-slope filters.

We propose to implement the programmable 16-slope equalizer with surface mount switches, resistors, inductors, and capacitors. This format is believed to have the best combination of desirable properties, including compactness, ease of assembly, availability of components, and low cost. The surface mount switches we propose to use are models HMC336MS8G and HMC321LP4 from Hittite Microwave Corporation. These devices have built-in digital decoders which simplify the control interface for the large switching network. They achieve -1.4 dB and -2.5 dB insertion loss at 4 GHz respectively, for a total insertion loss of 8 dB in the switch network. The worst case return loss for the switches is about -9 dB. Switches with better performance are available in die form if necessary, but we anticipate it will not be worth giving up the advantages of surface-mount components.

A tentative RF layout of the 16-step equalizer is shown in Figure 17. Of course, nothing about the layout constrains what range of slopes are actually used. The loss of the switches will skew the slopes toward the negative side, and of course other passive components in the system such as IF cabling may require that we have a greater range of positive slopes than negative ones. In any case, the layout shown occupies about 2.5 x 2.5 inches of board space. While the prototype equalizer would likely be connectorized for testing purposes, ultimately it makes sense to integrate it within the T304 Downconverter boards.

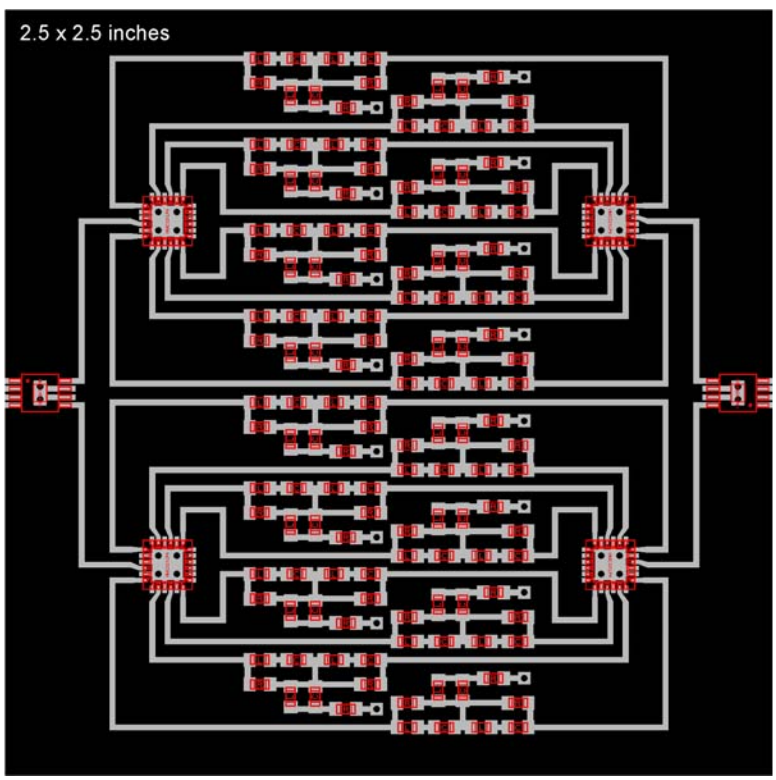

Figure 17 : Tentative RF layout for the proposed 16-Step Programmable Equalizer. Substrate is assumed to be 8-mils thick, Roger's 4003 Duroid, and the lumped elements are assumed to be in 0603 packages. Power supply and logic control lines are not shown. The circuit is shown at 1½ times larger than actual size.

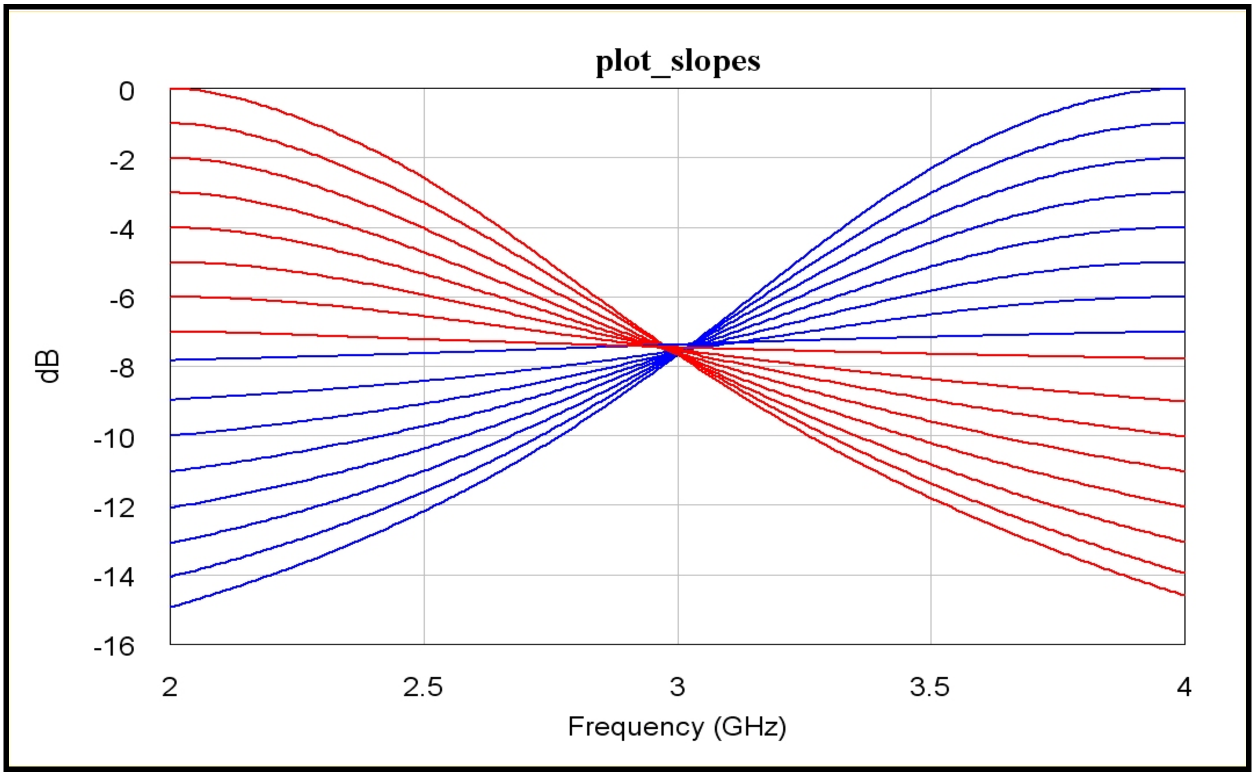

Figure 18 : Simulated performance of the proposed non-reflective filters with attenuators added to normalize the mid-band insertion loss.

Additional Options

As was described earlier, the different overall insertion loss between various states of the equalizer would make it necessary to adjust step-attenuators elsewhere in the system (ie: the Downconverter) to maintain a nearly optimal power level at the Samplers. Therefore, a potentially useful modification to the equalizer design presented in the previous section would be the inclusion of fixed, lumped-element attenuators in each of the rungs of the ladder. By choosing the proper attenuation in line with each sloped filter, the mid-band insertion loss could be made equal for all states of the equalizer, as shown in Figure 18. This would tend to decouple the optimum state of the step attenuators from that of slope equalizer. Obviously, the largest positive or negative slope used would limit the minimum mid-band insertion loss in this scenario. The return loss on one side would of course improve, and the phase characteristics would remain unchanged.

Alternatively, the attenuator values could be chosen to equalize the total integrated loss across the 2-4 GHz band, rather than the mid-band loss. Strictly speaking, this still would not guarantee that total input power to the Samplers is fixed, since the received signal itself could have structure in this bandwidth, particularly in the presence of RFI, but it may help to reduce the time required to search for the appropriate attenuator and equalizer settings during an observation.

One related option to consider is whether the equalizer should be made to look "transparent" to the system. This would mean the power level on the output would be maintained at roughly the same as the power level at the input, but obviously with a much flatter gain response. This would require the addition of a gain stage following the final output switch. As well as implementing the constant insertion loss scheme described above, which would require a total of 7.5 dB of loss (ie: half the maximum ±15dB gain slope) for each rung in the ladder, the extra loss in the switch network (about 8 dB) would have to be accounted for. An amplifier with ~15 dB would be adequate to bring the output power level back up. Many such surface mount amplifiers are available in the 2-4 GHz range with very flat frequency response. These amps are very small and typically cost around $2-$6, so this would represent a very simple and straightforward modification to the equalizer circuit.

It has been determined that the phase characteristics of the equalizer need not be "delay-like" for the EVLA, but this may not be the case for other telescopes. More linear phase responses may be achieved with relatively simple R and C networks, so long as the total slope is not too large. The high- and low-pass RC circuits shown in Figure 19 were optimized to get sloped responses up to -6 dB and +6 dB, respectively. The magnitude of insertion loss is shown in Figure 20. Unlike the earlier, more complicated structures, these circuits have greater loss in general, depending on the slope required, as well as poor return loss (less than 0.1 dB in some cases). The latter problem may be addressed by adding attenuators, or by using them in a balanced configuration with suitable quadrature hybrids. The primary advantage of these circuits is illustrated in Figure 21. It shows the amount of deviation from a best-fit linear ("delay-like") phase response for each of the slopes between -6 dB and +6 dB. The maximum phase error is roughly 1 degree over the 2-4 GHz band.

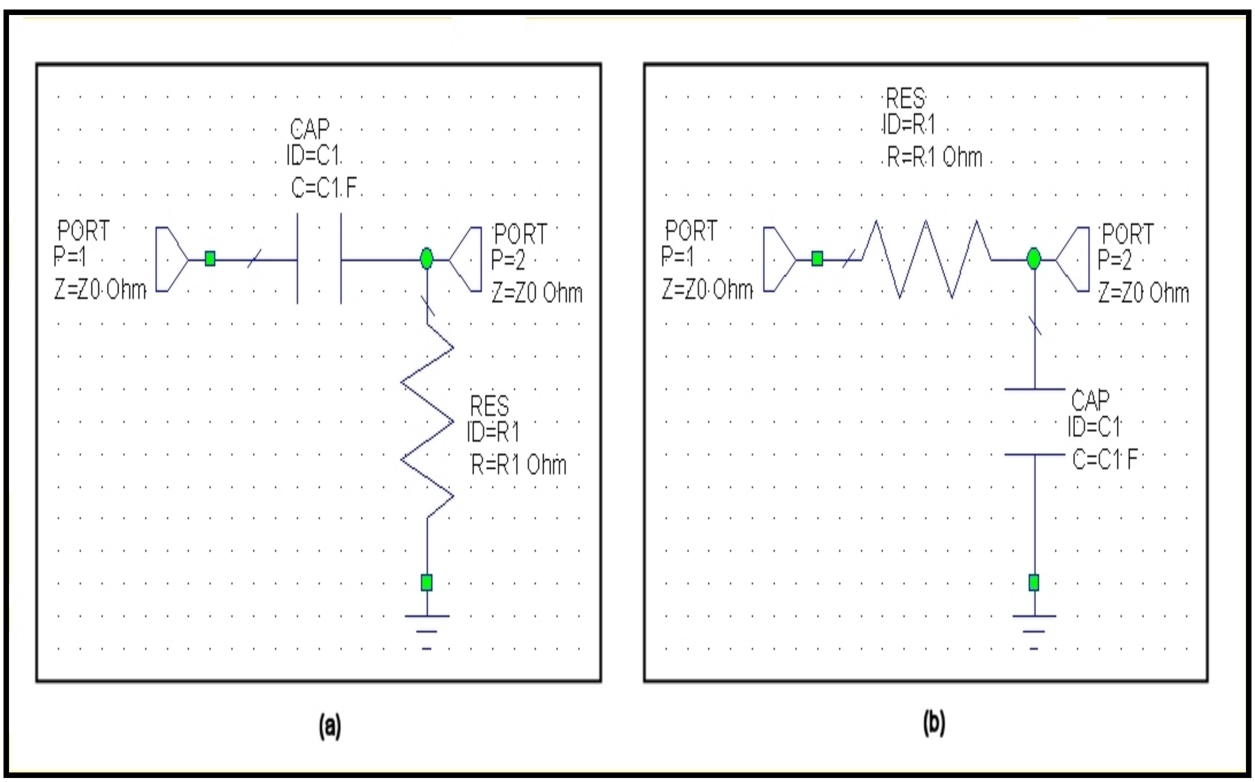

Figure 19 : Circuit topology for a) negative and b) positive slopes with delay-like phase characteristics.

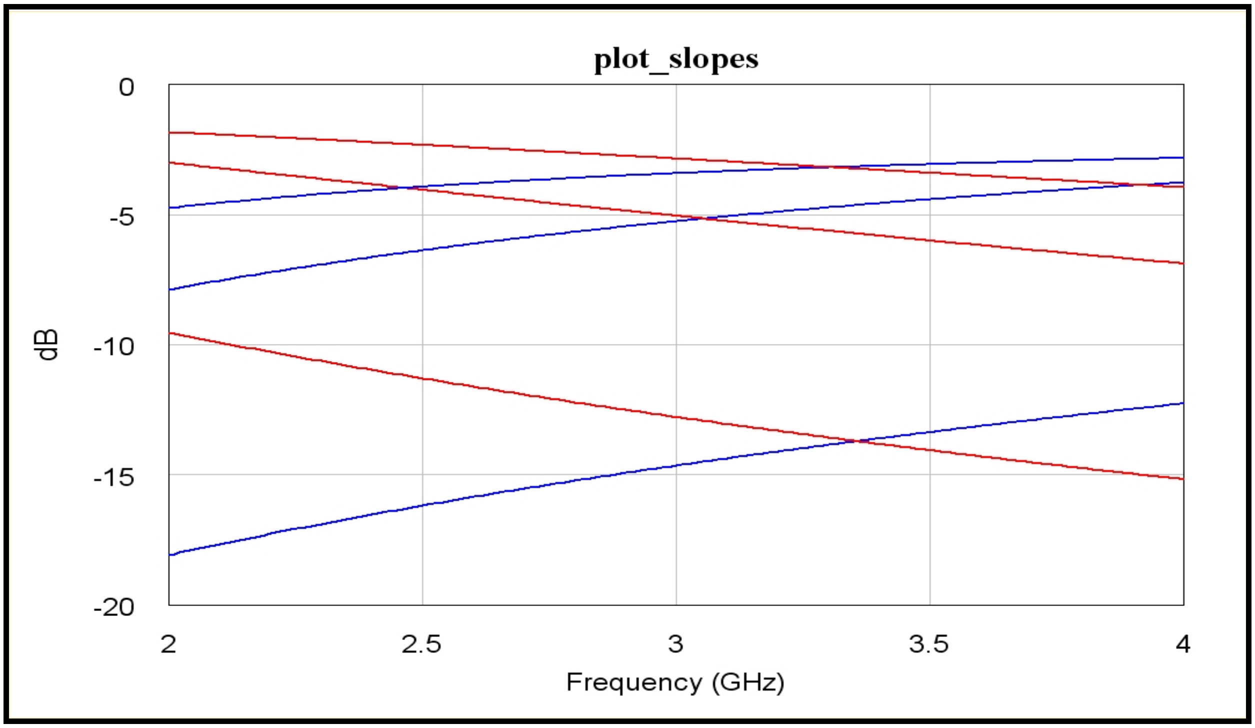

Figure 20 : Insertion loss of the RC circuits with linear phase, designed for slopes of ±2, ±4 and ±6 dB.

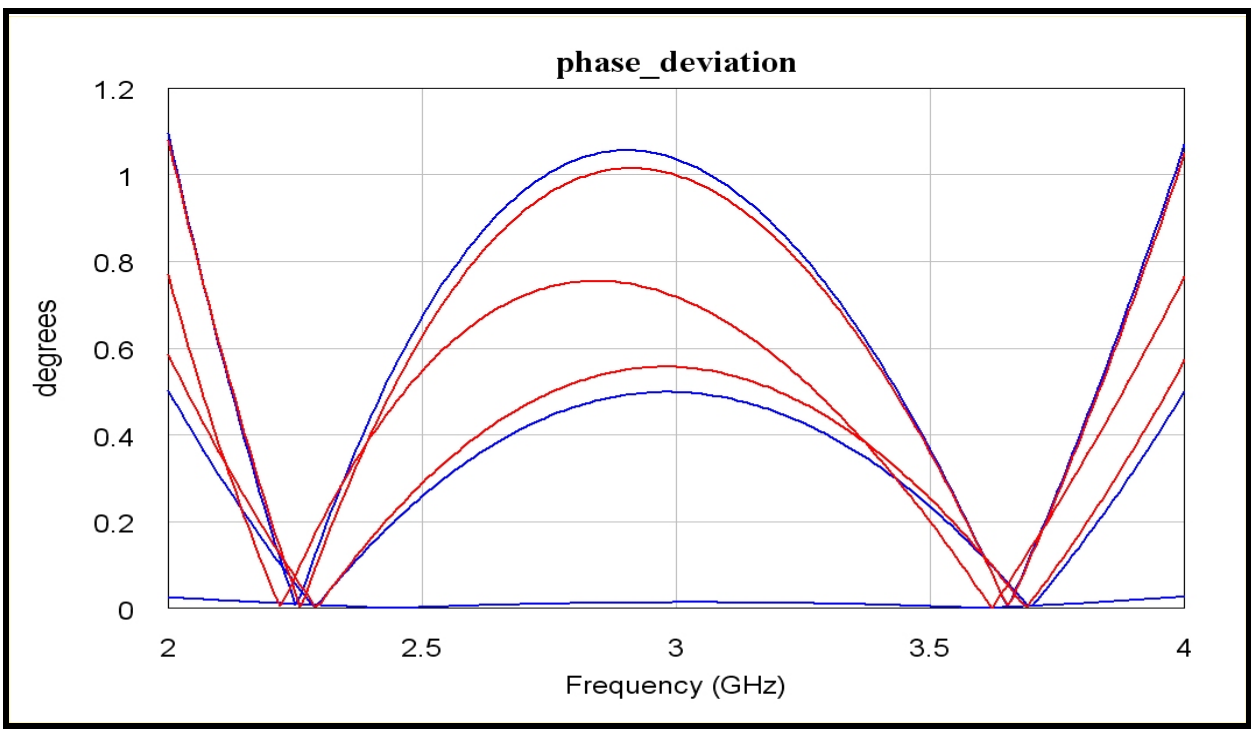

Figure 21 : Deviation from linear ("delay-like") phase for the RC circuits of Figures 19-20.

Conclusion

An equalizer will be required to compensate for the gain slopes arising in the RF and IF signal path. Unfortunately a single fixed gain slope equalizer would not be adequate since the gain slopes change in both magnitude and direction depending on which part of the RF band is being used. Additionally, they will be different from band to band and from receiver to receiver. Therefore, a 2-4 GHz programmable equalizer has been proposed that uses fixed non-reflective filters to provide selectable gain slopes. While a relatively simple equalizer with gain slopes between ±9 dB in 3 dB steps was suitable for correcting the bandpass of the K and Q-Band Front-Ends currently installed on the first EVLA Test Antenna, a much more capable scheme is recommended. The proposed equalizer would have 16 gain slopes settings between -15 dB and +15 dB in 2 dB increments. This unit would be not only be capable of compensating for the gain slope arising in the Front-Ends but the entire signal chain, including the Frequency Converters and cables, this making the 3 dB flatness design goal over any 2 GHz bandwidth within reach.

The design could be implemented using commercially available, surface-mount components, enabling the equalizer to be integrated with the microstrip IF boards already present in the system. It is estimated that the components needed for the proposed 16-step programmable equalizer would add less than $100 per channel to the cost of the T304 Downconverter. This would be increase the EVLA Project cost by about $24,000.

Among the disadvantages of this concept as it is presented is the fact that it only addresses the slope of the frequency response, and would not compensate for short-period ripple or suck-outs. However, the frequency response of the non-reflective lumped element structures used herein is quite flexible, and it is possible that they could be designed for concave-up or -down shapes with various depths in addition to the simple slopes presented earlier. The drawback of course is that you are limited in the total number of distinct equalizer curves that can be included on any single board. Different bands may require different R, L, and C values in the filters, but the overall design and board layout may be uniform.

Other drawbacks include the non-delay-like characteristic of the phase, and the overall insertion loss inherent in any passive equalization structure. The former problem may be alleviated by using simple RC networks in place of the non-reflective filters where the slopes are small. The latter problem can be addressed with the addition of normalizing attenuators in each ladder rungs and a 15 dB amplifier at the output of the equalizer.

Since the proposed equalizer has the ability to compensate for the accumulated gain slope of the entire receiver system, it may suggest that the tight EVLA flatness specification required of the receivers and frequency converters might be relaxed. It is obvious that the K and Q-Band Front-Ends will not meet the 2 dB spec at the band edges. However, since there is always a delicate balance between the noise added by the back-end amplifier stages and the overall headroom in the signal path, it makes sense to ensure that each sub-system is as flat as possible. That being the case, we would recommend that the gain flatness spec not be relaxed unless it requires significant extra cost to ensure compliance, or if the effort required to successfully obtain the flatness performance would result in an unacceptable delay to the Project. Finally, since the proposed equalizer will have little beneficial effect on gain ripple or suck-outs in the passband, great care must still be paid to the elimination of mismatches in the passband.

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