Frequency Domain Beamforming for a Deep Space Network Downlink Array

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Abstract— **This paper describes a frequency domain beamformer to array up to 8 antennas of NASA's Deep Space Network currently in development. The objective of this array is to replace and enhance the capability of the DSN 70m antennas with multiple 34m antennas for telemetry, navigation and radio science use. The array will coherently combine the entire 500 MHz of usable bandwidth available to DSN receivers. A frequency domain beamforming architecture was chosen over a time domain based architecture to handle the large signal bandwidth and efficiently perform delay and phase calibration. The antennas of the DSN are spaced far enough apart that random atmospheric and phase variations between antennas need to be calibrated out on an ongoing basis in real-time. The calibration is done using measurements obtained from a correlator. This DSN Downlink Array expands upon a proof of concept breadboard array built previously to develop the technology and will become an operational asset of the Deep Space Network. Design parameters for frequency channelization, array calibration and delay corrections will be presented as well a method to efficiently calibrate the array for both wide and narrow bandwidth telemetry.**

Table of Contents

1. Introduction

The ground portion of the telecommunication link between the Earth and NASA spacecraft engaged in solar system exploration is managed by the Deep Space Network (DSN). The DSN has three stations around the world in Goldstone, California; Madrid, Spain; and Canberra, Australia so as to always have at least one station in view of the target spacecraft as the Earth rotates. Each station has one large 70 meter dish antenna and multiple smaller 34 meter antennas that operate at S-Band (2 GHz), X-Band (8.4 GHz) and Ka-Band (32 GHz). These antennas are used for spacecraft navigation, tracking and telemetry.

Currently, additional 34 meter antennas are being added to the DSN network to augment the 70 meter antennas and to act as future replacements for the 70 meter antennas using arraying. Operational arraying capability has been provided for many years by the Full Spectrum Processor Array (FSPA) [1]. But, this system is nearing obsolesce, does not match the capability of current DSN telemetry receivers and offers no room for expansion. In a previous paper a breadboard array using a frequency domain beamforming architecture was described and the results of its performance were presented [2]. This paper presents plans for a new operational array, the Deep Space Communication Complex Downlink Array (DDA), to replace the current FSPA Array and describes the beamforming signal data path and key design parameters.

2. The DSCC Downlink Array

The DDA will be an eight input, one combined output beamforming antenna array. It will digitize IF inputs from 100 to 600 MHz with a 1280 MHz sampling clock, coherently combine 2 to 8 inputs digitally and transform the combined waveform back to analog at the same IF frequency. Compared to the older FSPA, it will handle twice as many inputs and processes 30 times more bandwidth (500 MHz versus 16 MHz). Although it uses frequency domain architecture for the signal data path, it still uses a full spectrum combining technique similar to the FSPA. The calibration algorithm will handle all types of telemetry and the DDA will accommodate all current data rates of DSN telemetry with capacity for future higher data rates. Calibration of the array allows the signals from each antenna to be coherently combined for an increase in signal to noise ratio of the telemetry data. Much of the calibration is performed by calculating the geometric delay of the incoming wave front to each antenna from a reference antenna position. However, tropospheric delays on antenna baselines as short as 200 m can vary significantly on time scales of only a few seconds, especially for Ka Band signals [3]. By obtaining cross-correlation measurements between

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antennas in the array, residual calibration data can be used to account for these variations and maintain the array in phase in real-time.

3. Time and Frequency Domain Beamforming

The DDA is required to process high bandwidth signals from multiple antennas simultaneously in order to coherently combine them in real time. Due to the RF to IF down-conversion in the DSN antenna front-end, part of the delay between antennas is converted to a phase offset. In addition, some antennas in the array are sufficiently spaced out that the rotation of the Earth causes a significant Doppler frequency shift from one antenna to another. So, the DDA must accommodate continuously varying delay, phase and frequency to keep the array in alignment.

Another design requirement is the ability to array telemetry bandwidths from 500 MHz all the way down to as low as 1 KHz. This implies a large amount of flexibility and configurability in the arraying and calibration architecture. It is especially challenging for antenna calibration, where the algorithms work best when the bandwidth being correlated closely matches that of the telemetry bandwidth.

While not a requirement, having a design that is easily scalable to a large number of antennas is a strong desire for the DDA to accommodate future expansion of the DSN. This pushes toward minimizing the amount of interconnects from one processing unit (board, chip, etc) to another.

At high sample rates, it becomes difficult to handle a digital signal as a single data stream in real-time processing devices such as Field Programmable Gate Arrays. These signals must be broken up into parallel data streams at lower clock rates. One way to do this is to break the original data stream up into multiple under-sampled time streams, each offset in phase by one sample. Alternatively, the signal can be broken up into multiple frequency bands.

In order to limit the combining loss due to delay errors, to less than 0.1 dB, the phase error across the band must be less than 20 degrees. The goal for the DDA is to be able to handle telemetry signals that occupy the entire 100-600 MHz band. Using a digital delay line, (e.g a FIFO), the best delay resolution possible is one sample delay at 1280 MHz, or 781 picoseconds. With this resolution, the maximum delay error possible for one input versus another input is half a sample and across the entire sampled bandwidth of 0 to 640 MHz this corresponds to a maximum phase error of +/- 45 degrees. So, in addition to the digital delay line, a fractional sample clock delay is necessary for the DDA.

When considering a full spectrum arraying architecture, two main classes of signal path architectures are possible, a time domain and frequency domain approach. Both have advantages and disadvantages, but can perform equivalent processing in different ways. For the DDA, a frequency domain approach was chosen to more efficiently handle large data rates in smaller frequency bands, and for the ability to perform the combining and calibration of the array in the same data path.

Time Domain

A time domain beamforming architecture performs all phase and delay corrections to the entire input signal at full bandwidth and sampling rate. Delays are performed with large sample rate digital delay lines and fractional delay

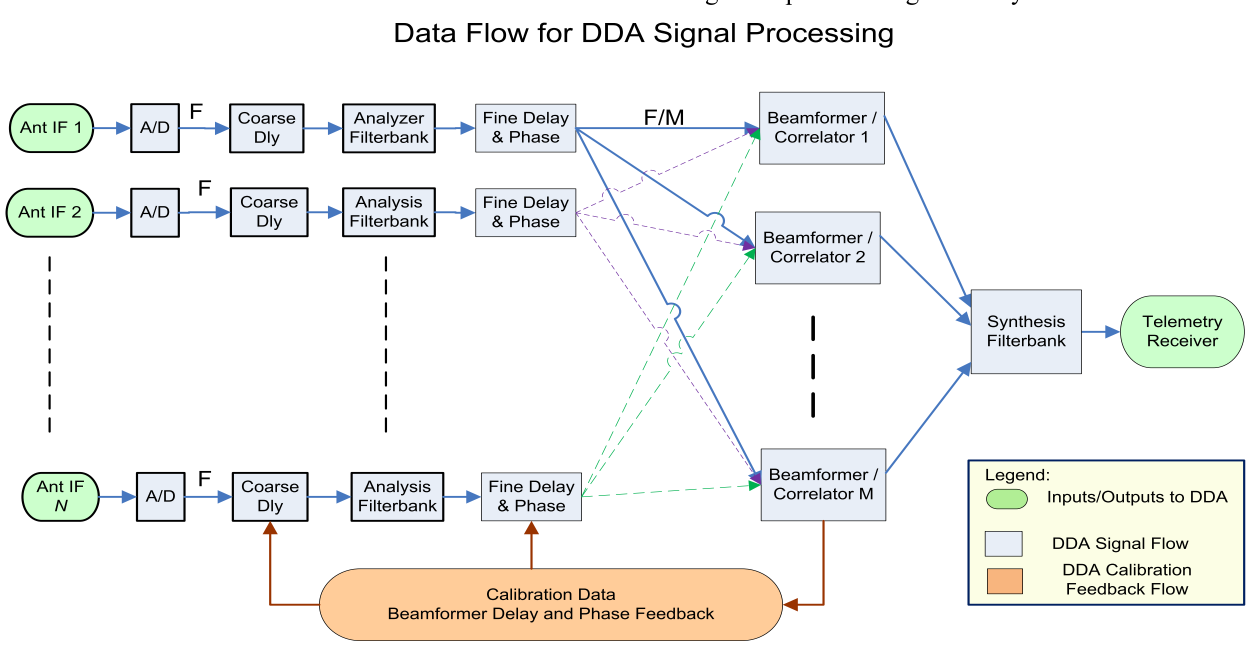

Figure 1: Frequency Domain Beamformer Data Flow

finite impulse response filters. Correlations for calibration are done off line and the input signal may need to be broken up into multiple smaller bands to improve correlation signal to noise ratio and to make delay measurements.

Frequency Domain

A frequency domain beamforming architecture breaks the input signal up into smaller frequency components. Each frequency component is combined separately and then resynthesized into a full bandwidth signal at the original sampling rate and bandwidth. Coarse delays can still be performed with a digital sample rate delay line at the original sampling rate, but fine, fractional sample rate delays are implemented as phase shifts to the individual frequency channels of the frequency domain beamformer. For wideband signals, the individual frequency channels are already in a format convenient for calibrating correlations. For narrow band signals, additional downconversion of lower rate frequency channels is less computationally intensive.

The main components of the DDA's implementation of a frequency domain beamformer are shown in Figure 1. After digitization, each digitized antenna data stream flows at rate, F. Coarse delay is applied using FIFO delay lines and an analyzer filterbank breams the time domain full bandwidth signal up into multiple frequency channels. Each channel receives individual phase and frequency corrections. Then then a portion of the channels are sent to M beamformer/correlator units. Each beamformer/correlator unit receives data from a given antenna at a rate of F/M. Both calibration and combining of the antennas take place in the beamformer/correlators for a 1/M fraction of the input bandwidth. Then, the synthesizer filterbank re-synthesizes the M different bands into one time domain combined output. Out of the main data path, once a second measurements of the correlation phase and amplitude allow a feedback algorithm to calculate updated phase and delay values to keep the array calibrated on the desired source in the sky.

4. ANALYZER AND SYNTHESIZER FILTERBANKS

The process of decomposing a signal into multiple channels, and then reconstructing the signal from those multiple channels is commonly performed with an analyzer-synthesizer filterbank pair.

One of the most common filterbank pairs is the Discrete Fourier Transform (DFT) filterbank. As shown in figure 2, the DFT filterbank uses complex exponentials to modulate the input, x(n), into K channels. Each channel is filtered by the low pass filter h(n), and downconverted by the factor M. The result is K channels equally spaced in frequency. Separate modifications to each channel can then be made, such as phase or amplitude adjustments. After these modifications, each channel is upconverted by M and passed through an anti-aliasing filter, f(n). Each channel is then modulated by another complex exponential to move it that channel back to its original frequency and all the channels are summed back to one time domain signal [4].

For the purposes of the DDA, a DFT filterbank pair has important and useful properties. A DFT filterbank breaks an input signal up into multiple equal frequency channels. With proper choice of filter coefficients, these channels can be perfectly or near perfectly reconstructed. And, with a DFT representation of the data, fractional clock delays can be implemented as a phase shift across the frequency channels.

A given channel in the analyzer filterbank is represented by

$$X_k(m) = \sum_{n=-\infty}^{\infty} h(mM-n)x(n)W_K^{-kn}, \quad k = 0,1,..K-1 \qquad (1)$$

where K is the number of frequency channels and M is decimation factor. Any modifications to these K channels yields a new set of signals, $\hat{X}_K(m)$. These channel signals are passed through a synthesizer filterbank to form the reconstructed signal as follows:

$$\hat{x}_n(n) = \sum_{m=-\infty}^{\infty} f(n-nM)\frac{1}{K}\sum_{k=0}^{K-1}\hat{X}_K(m)\,W_K^{kn} \qquad (2)$$

The analysis-synthesis filterbank used in the DDA is an oversampled (K/M > 1) discrete fourier transform (DFT) filterbank. Critically sampled filterbanks were not used for two main reasons. First, the aliasing errors due to the reconstructing the signals in the synthesis filterbank would have been too large [5]. Secondly, phase rate corrections done after the analysis filterbank independently on each filterbank would cause distortions in the synthesized signal and possibly place some carrier tones in the wrong location.

The filters h(n) and f(n) are designed to have the property of near perfect reconstruction. For near perfect reconstruction, the adjacent complex modulated filters of the filterbank must be approximately power complementary. A prototype filter meeting this criterion can be obtained by truncating the impulse response of a square-root raised cosine filter [6]. This prototype filter is used for both h(n) and f(n). The length of the filter should be a multiple of the number of channels, M, for an efficient implementation.

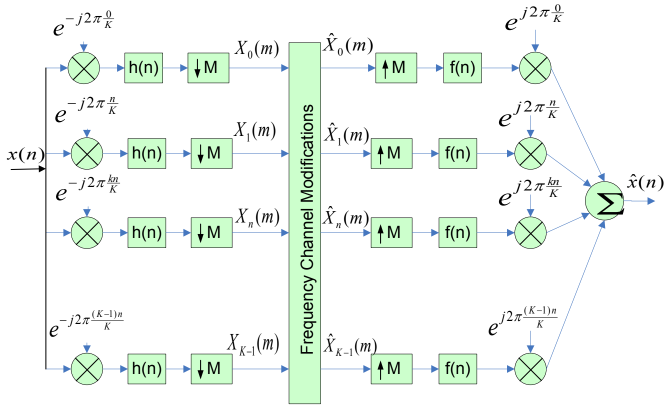

Figure 2: DFT Filterbank

Two other important factors in the design of the filterbank are the length of the prototype filter, h(n) and the oversampling ratio, K/M. The length of the filter determines the quality of the approximation of the square-root raised cosine characteristic. For actual implementation, the length of the filter should be a multiple of the number of channels, K. Acceptable results are obtained with filter lengths of 8K or greater. While critically sampled filterbanks are not satisfactory, the ratio K/M should be as small as possible to minimize data bandwidth when transferring channel data to correlators and combiners in the DDA. Matlab simulations of the filterbank with different ratios of K/M with K=512 showed that for a raised-cosine prototype filter of length 8K, an acceptable range for K/M was 1.2 to 1.5. A K/M ratio of less than 1.2 gave unacceptable distortion due to aliasing and ratios greater than 1.5 did not decrease distortion appreciably.

Since the acceptable range of K/M ratios is non-integer, a polyphase implementation of the DFT filterbank cannot be used. Instead, a Weighted overlap-add structure is used. This structure implements the filterbank in a block oriented manner instead of a polyphase implementation used with critically sampled DFT filterbanks.

The operations required in an analyzer filterbank are shown in Figure 2. Blocks of M samples are input into a bM shift register, where b=4 for this example. The shift register is then multiplied by a time-reversed version of the prototype filter, h(n). Then, this windowed sequence is divided into K blocks and the K blocks are overlapped and added together. This overlapping of K blocks is the reason why it is convenient to always have the length of h(n) be a multiple of K. The DFT of this overlapped sum is then taken. The result of the DFT represents a channelized output referenced to a sliding time scale, $r = n - mM$. Multiplying by the term, W_K^{-kmM}, changes the output to a fixed time scale, m=n/M.

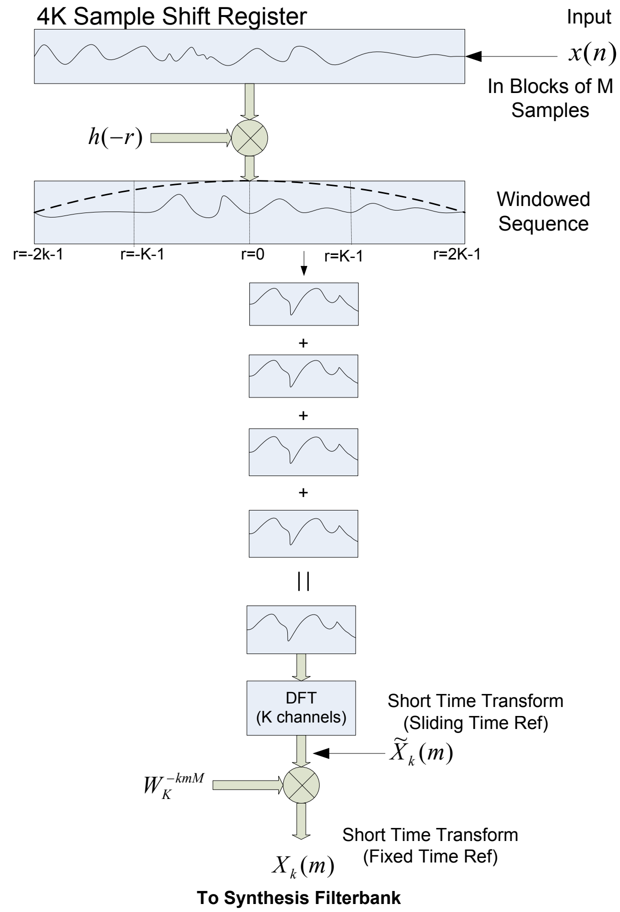

Figure 3: Analyzer Filterbank

The process of synthesizing the K channels back to one time domain signal is shown in Figure 3. First, the channels are transformed back to the sliding time scale by multiplying by W_K^{kmM}. Then, an inverse DFT is taken and b spectral copies are made to match the bK length of the prototype filter h(n). These spectral copies are then multiplied by the prototype filter and the output of this synthesis window is added to a bK length shift register. Similarly to input of the analyzer filterbank, blocks of M samples are shifted out at a time.

The last step of the analysis filterbank is a frequency rotation to change from a sliding time reference to fixed time reference due to the oversampling ratio. Similary, the first step of the synthesis filterbank is a fixed time reference to sliding time reference frequency rotation. These two steps can be skipped if the output of the analysis filterbank feeds directly into the synthesis filterbank. Similarly, the frequency rotation of the analysis filterbank is not necessary for cross-correlation of two frequency channels since frequency rotations would just cancel out.

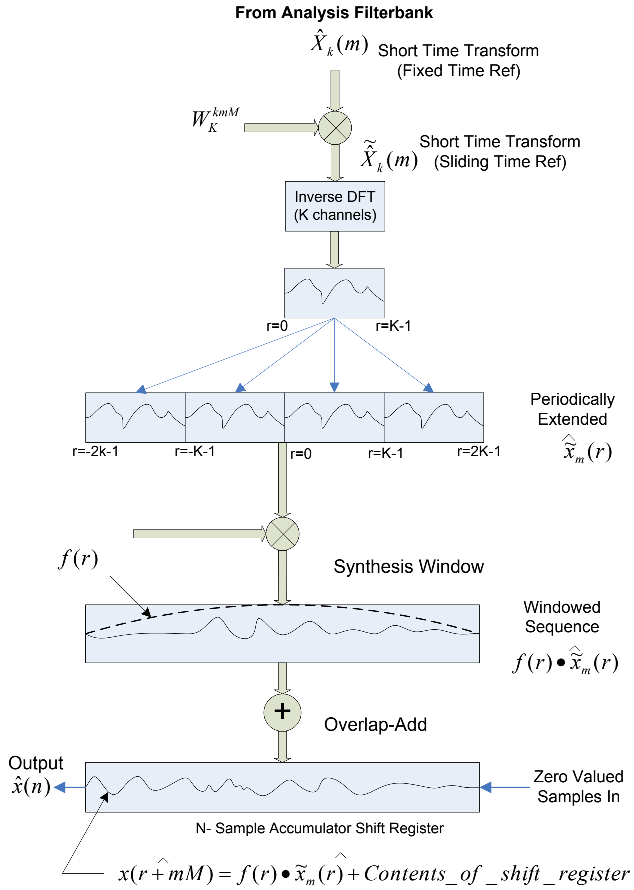

Figure 4: Synthesizer Filterbank

Later, when processing subchannels formed from the individual channels of the analysis filterbank, the fixed time to sliding time transformations must be taken into account when downconverting a portion of a channel to baseband.

5. Delay & Phase Corrections

For each antenna signal, the delay of the signal is adjusted by an integer sample clock delay line before the analysis filter bank and the phase and phase-rate are adjusted by a phase rotator following analysis filterbank. Using these adjustments, a local antenna of the array can be brought into phase and delay alignment with a reference antenna on a per channel basis. Figure 4 shows a simplified model of these delay and phase adjustments performed on a given channel. Using this model and a simple sinusoidal input signal, $S_1 = A(t)e^{\omega_{RF}t}$, the phase and phase rate equations for each frequency channel can be derived.

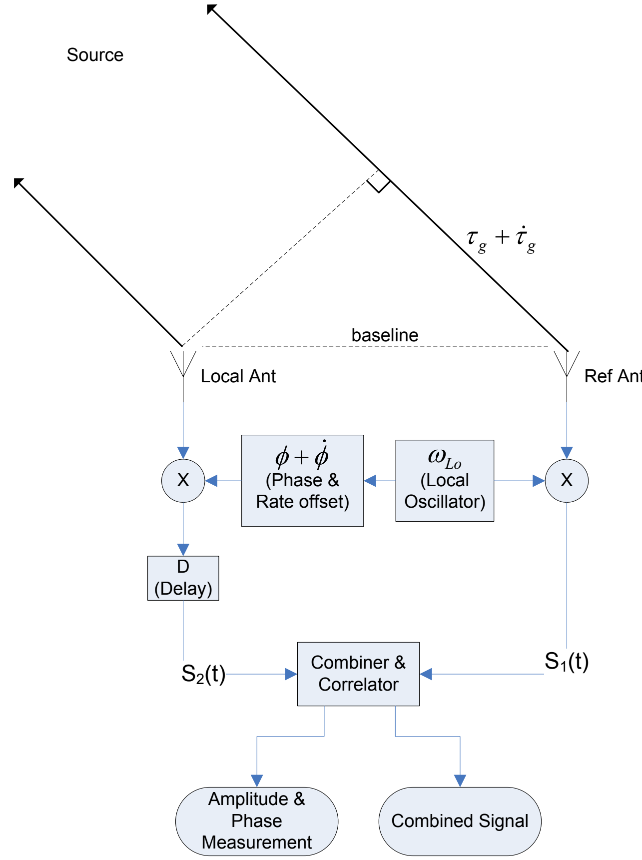

Figure 5: Simplified Model of phase and delay corrections for correlation and combining

Both the reference signal, $S_1(t)$, and local antenna signal, $S_2(t)$, are downconverted using a local oscillator. But, in addition, the local antenna signal has a phase and phase rate as well as a delay term, D, applied to align it with the reference signal.

The signals to be cross-correlated or combined can be expressed as,

$$S_1 = A(t - \tau_g - \dot{\tau_g} t)e^{\omega_{RF}(t-\tau_g-\dot{\tau_g} t)-\omega_{LO}t} \tag{1}$$

$$S_2 = A(t - D)e^{\omega_{RF}(t-D)-\omega_{LO}(t-D)+\dot{\phi}t+\phi} \tag{2}$$

In these equations, D represents the estimated delay with the resolution of a clock sample. The frequency of the local oscillator for converting from RF to IF is given by ω_{LO}, and ω_{RF} represents the center frequency of the input signal. The phase and phase rate can be determined by setting the phase terms equal and solving for ϕ, and setting the phase rate terms equal and solving for $\dot{\phi}$.

$$\phi = -\omega_{RF}(\tau_g - D) - \omega_{LO}D \tag{3}$$

$$\dot{\phi} = -\omega_{RF}\dot{\tau}_g \quad (4)$$

Since the delay, D, is implemented by a digital delay line with a resolution limited to the sampling period of the analog to digital converter, corrections to the phase must be made that are fractions of the clock period. These are, in fact, present in the phase equation, where the phase term, $-\omega_{RF}(\tau_g - D)$, represents a fractional clock delay correction. And, the phase term, $\omega_{LO}D$, represents a phase shift introduced by the RF to IF downconversion process. Finally, the term, $\omega_{RF}\dot{\tau}_g$, represents a Doppler shift in frequency due to the rotation of the Earth.

Phase and phase rate corrections are independently applied to each frequency channel output of the analysis filterbank. Each channel has it's own center ω_{RF} frequency. And by equation (1), the fractional difference between the clock delay and the geometric delay can be compensated for across the entire frequency band.

The residual phase at the edges of the 640 MHz complex sampling bandwidth due to the sample clock delay line is, at most, 0.25 cycles. Using the independent phase rotators for each channel, a channel will have a maximum phase error of 0.25/N cycles where N is the number of frequency channels across the entire band. For the 512 channel filterbank of the DDA, the maximum phase error of a channel is about 0.0005 cycles or about 0.18 degrees.

The oversampling of the analysis filterbanks allows phase rate corrections for geometric delay to be made in the channel specific phase rotators without distortion to the reconstructed signal in the synthesis filterbanks . As long as the phase-rate correction is less than ½ the extra bandwidth of a channel due to oversampling, the signal will still be in band, and not alias to the other side of the channel.

Each DDA analysis filterbank channel has 312.5 KHz excess bandwidth due to 1.25 times oversampling of each 1.25 MHz spaced channel. For the antenna geometry of the DDA, the largest baseline is about 20 km and the fastest delay rate of change is about 1.47 ns/s. Using equation(4) for a Ka Band input, the largest phase rate or frequency adjustment to be applied to each frequency channel output of the analysis filterbank is about 46 Hz. This phase-rate is well within the 156.25 KHz range for obtaining successful re-synthesis of the signal with the DDA synthesis filterbank.

6. ARRAY CALIBRATION

After the phase and phase-rate offsets are applied, the channelized IF signals are correlated and combined. Correlation is done on a frequency channel basis. Each frequency channel is cross-correlated against the corresponding frequency channel of all the other antennas in an array beam. The result will give an amplitude and phase measurement for each channel of a baseline. Also, to form an entire correlation matrix, auto-correlations are calculated. This correlation data can be used for interferometric analysis to calibrate fixed delays and antenna positions in the array and to calculate real-time delay and phase offsets to bring the antenna signals of an array beam into alignment for combining. The correlation measurements are dumped to a general purpose computer once a second. Delay, phase and amplitude corrections are calculated each second based on these correlation measurements. Several different algorithms have been implemented for deriving these corrections from the correlation data. They include SIMPLE, SUMPLE, EIGEN, and LSFIT, and are described in references [7] and [8]. SIMPLE uses the correlation of each antenna against a reference antenna, and SUMPLE uses the correlation of each antenna against the sum of the other antennas. Both of these algorithms make the calibration process an order N operation instead of an order N^2 process that uses the entire cross-correlation matrix. The remaining two algorithms make use of the entire cross-correlation matrix. For the DDA, the entire cross-correlation matrix data is formed in FPGA firmware. This is feasible for the 8 input DDA. But for future expansion to larger arrays, only the correlations of each antenna against the sum of the other antennas need be calculated. All of the algorithms also take advantage of the frequency domain nature of the array. It uses the frequency domain data to fit a telemetry amplitude profile across the band. In this way, interference for broadband sources such as planets can be nulled.[8]

Antenna signals are combined after the analysis filterbank and channel specific phase rotators. They are combined in the same FPGA that contains the correlators. Prior to the adder tree to combine the input signals for each channel, each input is multiplied by a scalar weight that is independent for each IF input. This is the heart of the integration of correlator and beamformer. Because the data are already broken up into frequency channels for correlation analysis, they can be combined on a frequency channel basis as well. The data needed for correlation are exactly the same that are needed for combining the frequency channels. Given the difficulty and expense of interconnects for these high bandwidth signals, processing correlations and combining in the same location provides a considerable savings. Further, the additional processing necessary to implement combining is a small fraction of that needed for correlations. For the correlator, the amount of processing is proportional to N^2 multiplies for a given bandwidth, where N is the number of antenna baselines. The combiner processing is proportional to N. Also, this integrated correlator and combiner architecture is flexible with regard to beamforming algorithms. Combining algorithms that require a full cross-correlation matrix can be accommodated as easily as those, such as SUMPLE, which only require a subset of a full cross-correlation matrix. The key element is that all the data for a given frequency channel is present in one place for all the antennas in an array

7. SUB-CHANNELS

The basic architecture of the DDA is optimized for arraying the full bandwidth available to a DSN channel (500 MHz). Calibration is performed by correlating the telemetry signal of one antenna against another antenna or the sum of the other antennas. The signal to noise ratio of the cross-correlation measurements is highly dependent on the bandwidth of the telemetry relative to the bandwidth of the frequency channels used in the analysis-synthesis filterbanks.

Telemetry bandwidths significantly smaller than the 1.25 MHz channels in the DDA suffer significant losses in correlation signal to noise ratio that degrade the ability of the array to lock up to a signal and obtain combining gain.

This can be demonstrated by considering signals to be correlated in single channel.

$$s1(t) = d(t)e^{j\emptyset 1} + n1(t)$$

$$s2(t) = d(t)e^{j\emptyset 2} + n2(t)$$

Where each signal contains a common telemetry component d(t), each with its own phase offset. Also, assume the delay offset between the telemetry signal component in each signal is negligible. And, each signal has an independent white Gaussian noise component.

The noise power in both signals is given by $N_{01}B_C$ and $N_{02}B_C$, where B_C is bandwidth of the channel and N_{01} and N_{02} are the power spectral densities of the noise in each signal. For a band-limited telemetry signal, its power can be roughly estimated as P_dB_d, where P_d power spectral density of the telemetry and B_d is the bandwidth of the telemetry.

After integrating the cross-correlation outputs to a bandwidth of B_{ACC}, the signal to noise ratio of the cross-correlation is given by,

$$C_{SNR} = \frac{P_d^2 B_d^2}{N_{o1} N_{o2} B_c^2}\left(\frac{B_c}{B_{ACC}}\right)$$

The telemetry bandwidth will be some fraction of the noise bandwidth $B_d = \alpha B_c$, where α is between 0 and 1. Then, the cross-correlation SNR reduces to,

$$C_{SNR} = \frac{P_d^2 \alpha^2}{N_{o1} N_{o2}}\left(\frac{B_c}{B_{ACC}}\right)$$

The cross-correlation SNR is maximized when the telemetry bandwidth in the channel matches the noise bandwidth in the channel at $\alpha = 1$. In the DDA, frequency channels are a fixed bandwidth. But for low rate telemetry smaller bandwidth channels are desired. For this reason, the DDA forms sub-channels.

The DDA will have a number of sub-channels chosen from the available band of 512 channels. Each sub-channel will have its own mixer to move a region of interest to baseband followed by simple accumulate and dump decimation filter. The sub-channels are like a second stage analysis filterbank, only more flexible. The number of sub-channels will stay fixed, but the bandwidth of the sub-channels will be adjustable from 100 KHz down to 125 Hz. With each sub-channel being completely independent, they do not share the efficiencies of the DFT filterbank implementation of the 512 channels. However, they each run at 1/512 the rate of the filterbank processing, so such efficiency is not essential.

Sub-channels also serve a second purpose. The calibration algorithms use the information in cross-correlations of multiple frequency channels to fit an expected profile of the telemetry spectrum. Among the several parameters defining the spectrum is its position in frequency. To derive this position accurately, ideally at least 4 or more channels are required by the calibration algorithms across the telemetry width. So, for telemetry bandwidths less than 5 MHz, sub-channels are desired.

In the feedback processing, either channel or sub-channel information can be used to match to a telemetry spectrum. For narrow band telemetry, however, sub-channels are more flexible than simply forming a second stage of filterbanks after the initial 512 channel filterbank. First, the sub-channel bandwidths are flexible. This could be done in a DFT filterbank, but adds significant complexity in hardware. Second, the sub-channels need not be contiguous. Non-contiguous sub-channels could not be done in DFT filterbank. The non-contiguous sub-channels are useful since many low rate telemetry signals modulate telemetry on a sub-carrier modulated on a carrier. This gives upper and lower sideband components and a carrier component that are not contiguous in frequency. With the 64 sub-channels, some portion of the sub-channels can be tuned to cover all three of these components.

8. SUMMARY

A frequency domain beamforming architecture is planned for the implementation of a downlink arraying system for the antennas of DSN. Compared to current arraying equipment, it increases the available arrayed bandwidth from 16 MHz to 500 MHz, doubles the number of antennas that can be arrayed and uses an architecture with room for future expansion in the number of antennas.

The architecture was developed and characterized for a DSN Array Breadboard. Current efforts are underway to build upon that design and turn it into an operational asset of the DSN. One key enhancement is the addition of sub-channels to handle narrow band telemetry effectively.

9. ACKNOWLEDGEMENTS

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BIOGRAPHIES

Robert Navarro *received a B.S in Engineering from Harvey Mudd College in 1986 and a M.S. in Electrical Engineering in 1987. He has been with JPL for more than 19 years. He has worked on DSN ground systems for radio science and VLBI recording as well as correlators and antenna arrays as an engineer and a manager. He is currently the supervisor of the Processor Systems Development Group. His career started with Hewlett Packard working on circuit design and image processing for ink jet printers.*