Self-Interference Cancellation in Full Duplex using Padė Approximation

A project report submitted by

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1. Introduction

Today's mobile users want faster data speeds and more reliable service. The next generation of wireless networks 5G, promises to deliver that, and much more. 5G wireless network would be able to handle 1000 times more traffic than the existing ones. It will be 10 times faster than the current 4G LTE. This technology would be a foundation for the development of new technologies like Autonomous Vehicles, Virtual Reality, Internet of Things and many more.

To achieve this, wireless engineers are designing a suite of brand-new technologies. Together, these technologies will deliver data with less than a millisecond of delay (compared to about 70 ms on today's 4G networks) and bring peak download speeds of 20 gigabits per second (compared to 1 Gbps on 4G) to users. If all goes well, telecommunications companies hope to debut the first commercial 5G networks in the early 2020s.

Figure 1: A figure showing the broad advantages of 5G

At the moment, it is not clear of which technologies exactly go into the 5G network. But some of the trending topics regarding 5G are millimeter waves, smart cells, massive MIMO, Beam forming and Full Duplex. This project mainly concentrates on 5G Full Duplex and the challenges present in achieving that.

Full Duplex:

Today's base stations and cellphones rely on transceivers that must take turns if transmitting and receiving information over the same frequency, or operate on different frequencies if a user wishes to transmit and receive information at the same time.

With 5G, a transceiver will be able to transmit and receive data at the same time, on the same frequency. This technology is known as full duplex, and it could double the capacity of wireless networks at their most fundamental physical layer: Picture two people talking at the same time but still able to understand one another—which means their conversation could take half as long and their next discussion could start sooner.

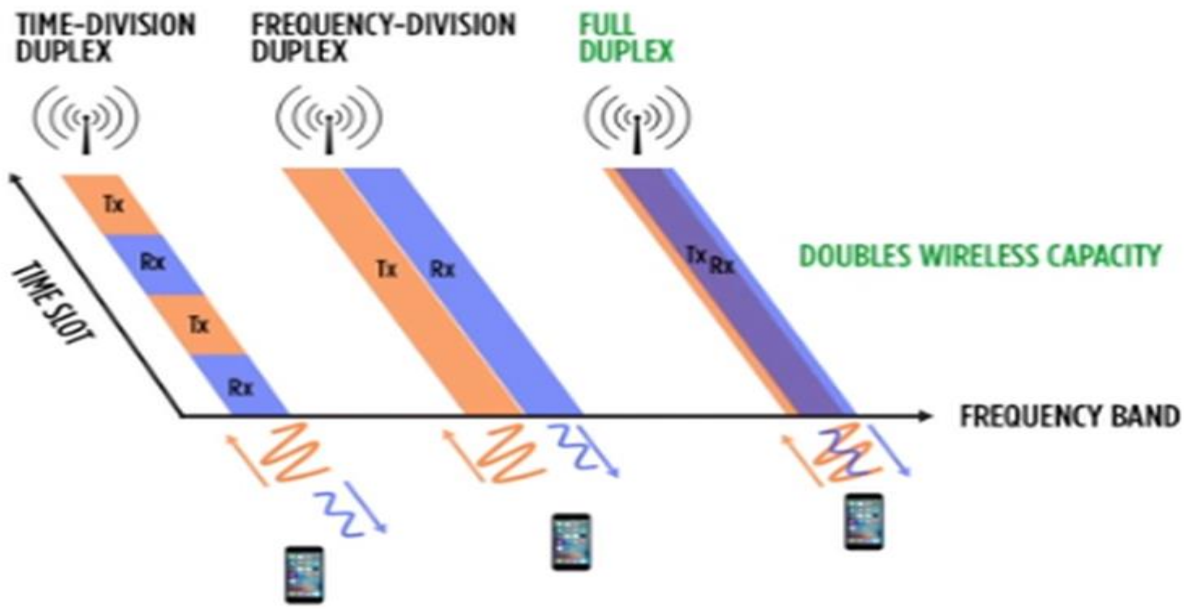


Figure 2: A figure showing full duplex wireless communication

Advantages of Full Duplex:

- ***Effectively doubles spectrum efficiency:*** By employing a 5G full duplex scheme, only one channel is needed to transmit data to and from the base station rather than two for an FDD scheme, or when using a TDD scheme the full transmission time can be utilized in both direction rather than half - the scheme effectively makes TDD schemes redundant. This effectively doubles the spectrum efficiency.

- ***Fading characteristics:*** As the same channel is used in both directions the fading / propagation characteristics will be the same. Difficulties can arise using an FDD scheme when one channel is affected by fading and the other is not affected. This issue doesn't exist now.
- ***Filtering:*** FDD requires filters to be used to ensure that the transmitted signal does not enter the receiver and desensitize it. As more bands were added, more filters were required with a resulting increase in loss and drop in performance. By using single channel 5G full duplex, this issue can be overcome as techniques used have been shown to be consistent over a wide bandwidth.
- ***Novel relay solutions:*** The techniques used for 5G full duplex on a single channel enable the simultaneous re-use of spectrum in backhaul as well as the main user access can allow for almost instantaneous retransmission.
- ***Enhanced interference coordination:*** The simultaneous reception of feedback information while transmitting data, possible using 5G full duplex reduce the air interface delays and provide much tighter time / phase synchronization for techniques like Coordinated Multipoint, CoMP etc

Challenges in Full Duplex:

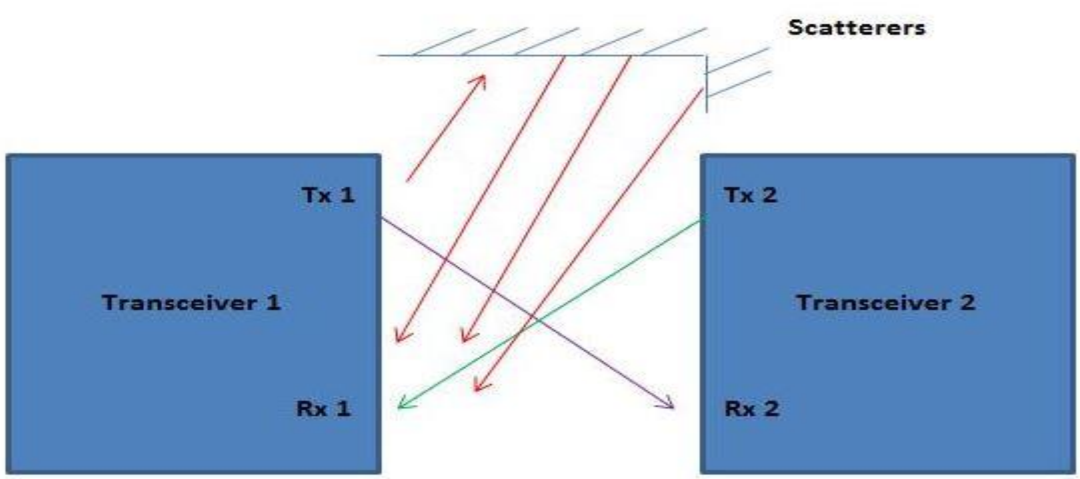


Figure 3: Figure showing the Self Interference in full duplex

The above figure clearly explains the biggest challenge present in achieving Full Duplex which is the Self Interference. So, the major task in achieving Full Duplex is to cancel this Self Interference at different stages. A cancellation of at least 110 dB is needed all together to suppress the Self Interference signal to the noise floor. The major of these cancellation stages would be the following:

Electrical balance isolation: The isolation technique employed effectively uses the same technology as used in landline telephones to provide isolation between the incoming and outgoing signals and this is obviously modified for RF. It can provide around 20dB of isolation

At the level of analog circuits (before the A/D converter): This implements self-interference cancellation in the analog domain using a noise cancellation circuit. The transmit signal is fed to the circuit as a noise reference, which subtracts it from the received signal, after adjusting for phase and amplitude.

In the digital baseband: This uses the received digital samples after the analog-to-digital conversion in the receive path. The transmitted samples are stored in a local memory. The received samples are correlated with the transmitted samples to determine the beginning of the transmitted packet and its phase in the received samples. The transmitted samples are rotated samples that almost completely remove the transmitted signal from the received signal.

RF self-interference cancellation: The main amount of reduction of the transmitted signal is provided by using RF cancellation techniques - often referred to as self-interference cancellation, SIC. Much investigation work has been ongoing to improve the performance and enable 5G full duplex in a single channel to be a realistic option.

The main goal of this project is to focus on the RF self-interference cancellation and this is attempted to achieve using Padė approximation. Section 2 gives a brief picture on the relevant theory for the project. Section 3 and 4 focuses on the problem statement and simulations done for validation of the idea. Section 5 explains how these filters can be built practically and concludes the discussion by mentioning the scope for the future work.

2. Theory

Basic Concepts of Wireless Communication:

Carrier Modulation:

Consider a complex baseband signal $x(t)$ with bandwidth f_s and carrier signal $c(t)$ with a frequency f_c.

Generally $c(t)$ will be just a sinusoid of frequency f_c and a phase shift $\emptyset$.

$$c(t) = \exp(j2\pi f_c t + \emptyset) \qquad (1)$$

Modulation is a technique where a low frequency signal is multiplied with a high frequency carrier for the purpose of transmission.

So, the modulated signal in our case would $s(t) = x(t) * c(t)$

$$\text{Hence, } s(t) = x(t)\exp(j2\pi f_c t + \emptyset) \qquad (2)$$

The above representation of $s(t)$ is often called complex baseband representation.

Multipath:

A transmitted signal $s(t)$ before reaching the receiver goes through various losses. At a large scale, path loss due to various structures and environment and shadowing due to very tall buildings occur. But, when we come to a small scale level i.e. within 100 λ, (λ is the wavelength of the carrier) we see multipath fading as a major component of signal distortion.

This happens because of the various reflectors and scatterers present around the antenna. If we fix a time resolution, all those scattered components that come in that interval are seen as a single component. These components actually vary in magnitude with the time.

So, multipath looks like an impulse response in the delay domain but this impulse response varies with the time. This type of variation of the multipath with time and delay is characterized by a power-delay profile.

An example of a multipath where different taps are plotted against delay and time is shown in the figure below.

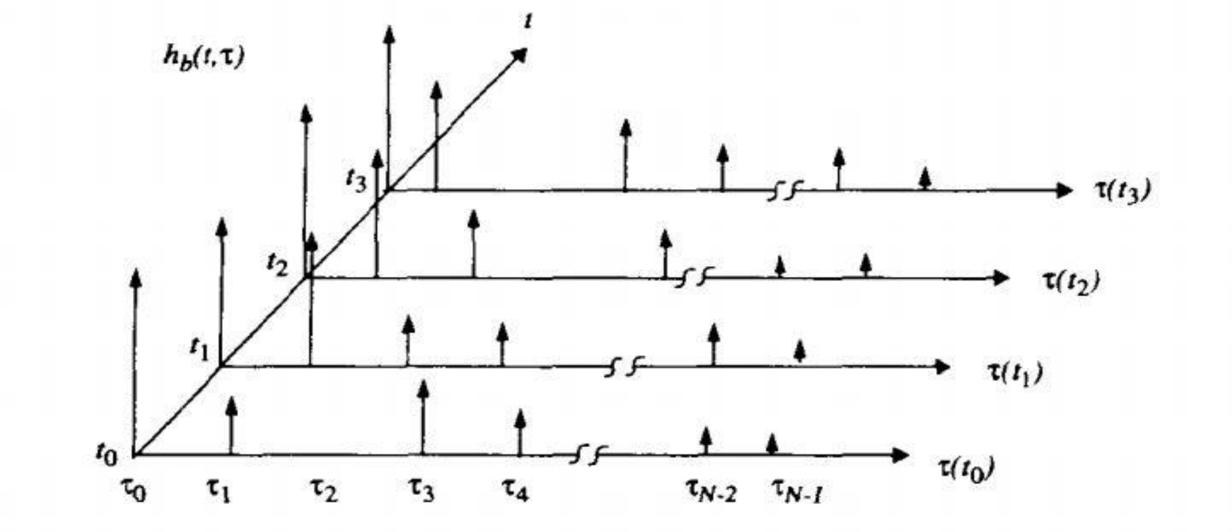


Figure 4: A figure showing the multipath variation with delay and time

A mathematical expression for the multipath that is considered for the project is based on the following interference model.

Interference Model:

The interference model is as follows:

- The Transmitted signal undergoes multiple reflections with the surroundings before it reaches the receiver end
- Based on the path these components travel, we have delays that are proportional to those distances and amplitude attenuation which is inversely proportional to the distance
- Now, all these components get added to form the final Interference signal
- So, the unknowns are the no of reflected paths, their delays and amplitudes
- But, for the purpose of analysis, some control over the variations of these unknowns is assumed and the cancellation achieved in each case is observed

For k multiple reflections, the interference signal is as below

$$I(t) = \sum_{i=1}^{k} \alpha_i x(t - \tau_i) \qquad (3)$$

Here α_i is the amplitude and τ_i is the delay of the i^{th} reflected path of the signal.

Taylor Approximation & Padẻ Approximation:

The transfer function of an ideal delay element is e^{-st_d}. But, it is practically not possible to implement this on a circuit with lumped elements. In order to implement this transfer function using lumped elements, Taylor approximation is used to find an nth order polynomial which approximates the delay.

$$e^{-st_d} \approx 1 - t_d s + \frac{{t_d}^2}{2!} s^2 - \frac{{t_d}^3}{3!} s^3 + \cdots + \frac{(-t_d)^n}{n!} s^n \qquad (4)$$

Since, this is only a numerator polynomial, it is still not realizable. A denominator polynomial of order m (≥n) is needed to make it realizable. This D(s) should have a relatively constant magnitude over the frequency range of interest. Thus, the modified $H_d(s)$ is as below

$$H_d(s) = \frac{1 - t_d s + \frac{{t_d}^2}{2!} s^2 - \frac{{t_d}^3}{3!} s^3 + \cdots + \frac{(-t_d)^m}{m!} s^m}{D(s)} \approx e^{-s(t_d + t_f)} \qquad (5)$$

Coming to the implementation of $H_d(s)$, $\frac{1}{D(s)}$ is realized in the first stage and $N(s)$ in the second stage. The block diagram of this is shown below in Figure 5

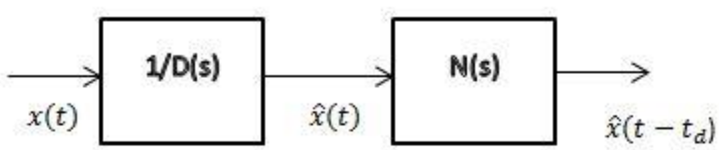


Figure 5: A figure showing the implementation of Hd(s)

The first stage is implemented as a Bessel filter with a cutoff frequency chosen so as to maintain a constant magnitude over the entire bandwidth of the signal. This results to an output signal which will have the same amplitude of the input signal and a constant group delay t_f. So, the output of the first stage is actually just a delayed version of $x(t)$ and hence, $\hat{x}(t) = x(t - t_f)$. Coming to the second stage, if the coefficients are chosen appropriately as per the Taylor approximation mentioned above, the final output can be viewed as $x(t - t_f - t_d)$.

Tunable delays for the input signal are achieved by varying the coefficients of N(s). For $H_d(s)$ to look like e^{-st_d} , N(s) is chosen based on Padė approximation.

Padė approximation gives the best approximation of a function as a rational function of the same order.

Given a function f(x) and two integers m≥0, n≥1; a padė approximant R(x) is given by:

$$R(x) = \frac{\sum_{j=0}^{m} a_j x^j}{1+\sum_{k=1}^{n} b_k x^k} \qquad (6)$$

Here, all the derivatives up to order (m+n) evaluated at 0 should be same for f(x) and R(x).

In our case, we first consider a truncated polynomial approximation for e^{-st_d} and find a rational function $N(s)/D(s)$ such that

$$N(s) = \left(1 - s\left(t_d + t_f\right) + s^2\left(\frac{(t_d+t_f)^2}{2!}\right) - s^3\left(\frac{(t_d+t_f)^3}{3!}\right) + \cdots\right) D(s) \quad (7)$$

In a conventional Padė approximation problem, both N(s) and D(s) can be varied until the above equation is satisfied. But in this case D(s) is fixed and coefficients of N(s) are varied. But, in our case D(s) is kept fixed and the coefficients of N(s) are varied.

In this way signals can be generated which are amplitude scaled and delayed versions of the original signal. So, these signal outputs form a key role in the case of self-interference cancellation in our project.

OFDM Waveform Generation:

OFDM signals are generated in practical cases using the following architecture in the figure below. The figure shows a complete OFDM system using DFT based architecture.

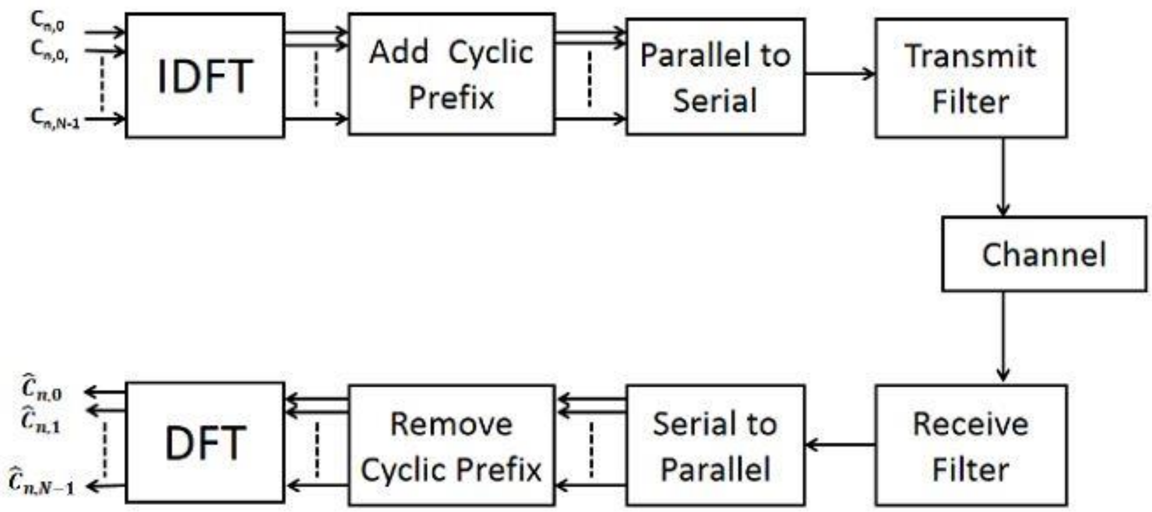


Figure 6: A figure showing the implementation of an OFDM system

The procedure is as explained here:

- First collect the bits received pertaining to a single block/symbol
- Find the IDFT for the bits that are collected above
- Add cyclic prefix for the output
- Then, convert the bits back to a serial stream

This gives the OFDM waveform for 1 symbol. Repeat the same for the total number of symbols that are present. The idea that every coefficient of an IDFT matrix is orthogonal makes the signal frequencies orthogonal.

This OFDM signal is now transmitted across the channel. At the receiver, convert the serial data again to parallel and then remove the inserted cyclic prefix. Now, take the DFT of the signal to get back the data bits corresponding to each symbol.

3. Problem Statement

Existing Circuit Explanation:

In the case of a single antenna system, a shared architecture with a single antenna is used for both transmission and reception. As mentioned already, a cancellation of at least 110 dB is needed. Taylor Approximation is used to estimate the interference signal.

An isolator is present in the antenna which provides certain amount of initial cancellation. Then, Cancellation in RF domain is to be performed before the LNA in the below pic.

This design has a vector modulator which takes the transmitted signal as the input, alters its amplitude and phase in order to match the direct signal component in the interference signal. This design thus accommodates for the cancellation of the signal component alone in the RF domain.

It doesn’t allow any cancellation of the derivatives and higher order components. So, the cancellation achieved will be limited.

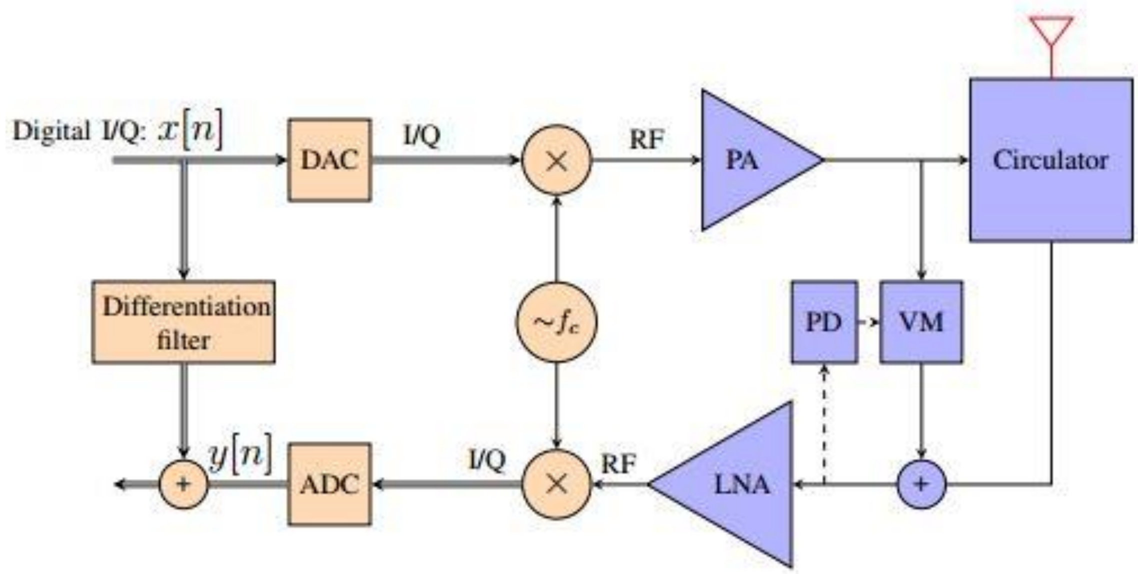


Figure 7: This figure taken from the reference [4] just for illustrative purpose of the circuit

The results for cancellation using Taylor approximation are mentioned in the next page.

The circulator used in this case provides 18 dB of cancellation. The RF/analog cancellation reported for different cases:

For 10 MHz 4-QAM signal modulated on a single carrier with Tx power = 4dbm, the cancellation achieved is 57 dB (inclusive of circulator cancellation)

For 20 MHz OFDM signal with multiple subcarriers and Tx power = 4dbm, the cancellation achieved is 54 dB (inclusive of circulator cancellation)

A plot showing the power spectrum of signal before & after cancellation is below.

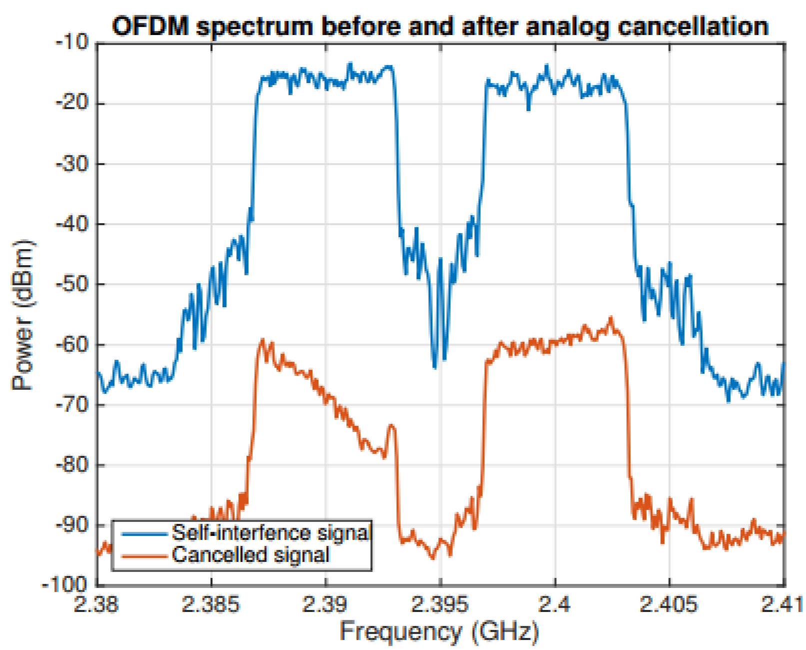


Figure 8: This figure shows the power spectrum of signal before and after cancellation

Drawbacks for this model:

This model works fine only when the delays of the interference signal actually fall in the range where Taylor approximation holds good. But, Taylor approximation is valid only for small values of delays. The range is defined as the maximum delay that can be achieved for the signal with minimal distortion. For a monopulse (t_p = 35ps) , the maximum range of delay achieved as per above statement is $|t_d \leq 40$ ps$|$.

Because of the above limitations, the cancellation achieved using Taylor approximation is limited. For higher cancellations, a good approximation for the greater values of delays is required. The padė approximation gives a better approximation for higher values of delays than the Taylor approximation. For a monopulse (t_p = 35ps) , the maximum range of delay achieved as per padė approximation is |$t_d \leq$ 70 ps|.

So, this project exploits the padė approximation in designing the tunable filters which can give higher RF/analog cancellations.

Problem Statement:

So, as far as this project is concerned, the vector modulator portion is to be replaced with a tunable filter. The interference is assumed to follow the model as already mentioned. So, an equivalent block diagram of the project is as below

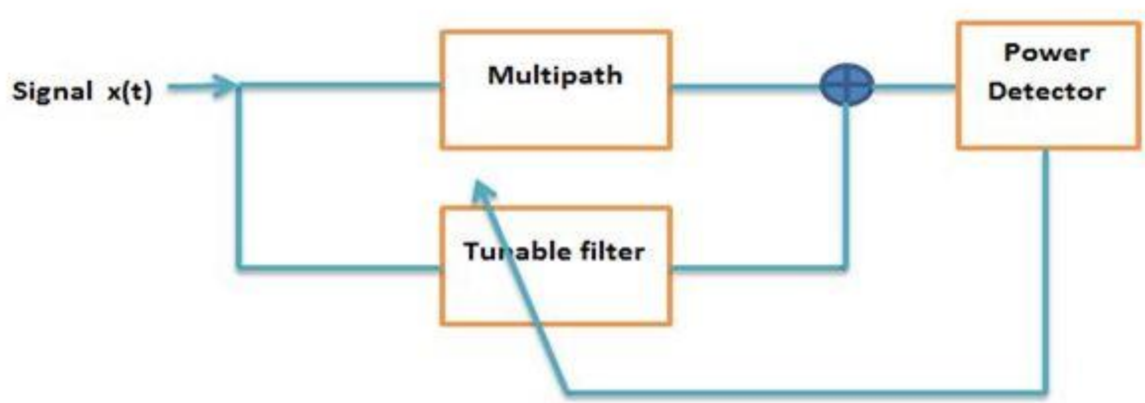


Figure 9: A figure showing the block diagram of the project

Consider the impulse response of the tunable filter to be $h(t)$ and the interference signal as in equation 3,

The output of the tunable filter is, $R(t) = x(t) * h(t)$

The cancelled signal after the adder is, $e(t) = I(t) + R(t)$

The power of $e(t)$ as seen by the power detector is used as the reference to tune the coefficients of the filter. And finally a minimum power for $e(t)$ is achieved which is analogous to the maximum cancellation achieved for that interference environment. So, the next question is how to build a tunable filter. This question is addressed in the next section.

4. Simulations

Refer to the block diagram in Figure 3. This is the structure used for the entire analysis.

Different initial approaches were tried out for the implementation of these tunable filters. All the approaches have 3 main parts : signal generation, interference generation and filter response. Different approaches differ in at least one of the 3 parts mentioned.

Approach 1:

Signal Generation: $x(t)$ is a two tone sinusoid with center frequency around few MHz and baseband signals at around few kHz. This is a scale down version of signals with MHz's bandwidth and GHz's of center frequency. The scale down is for the convenience to work with low complexity and less data sizes.

Consider a particular example where carrier frequency = 2 MHz and the tones are at 1kHz on the either side of the carrier.

The time domain and frequency domain plot of the signal is as below:

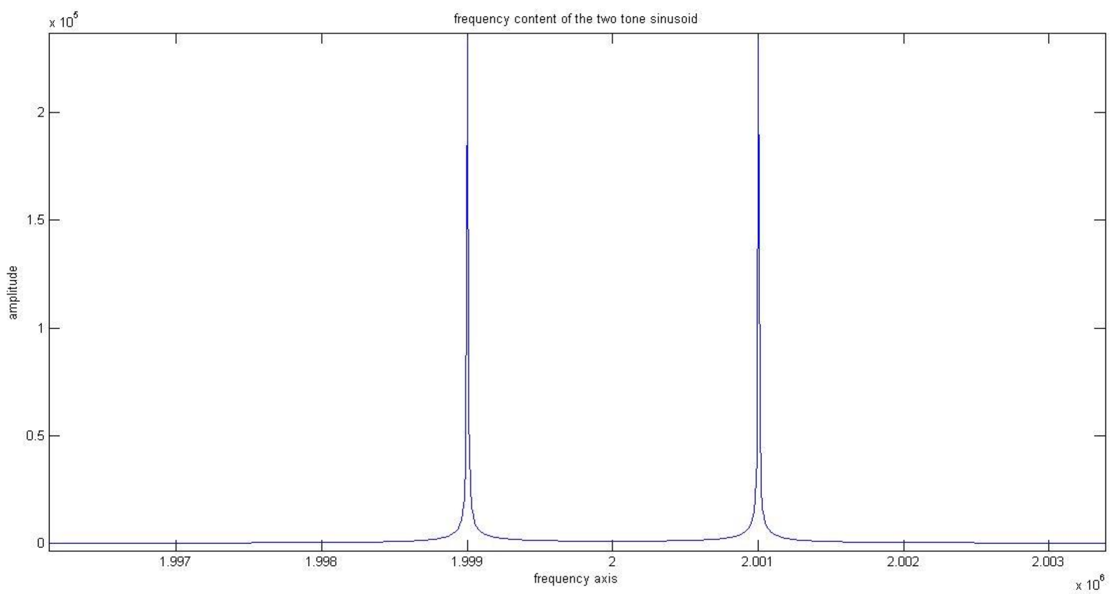


Figure 10 : Frequency Response of a two tone sinusoid

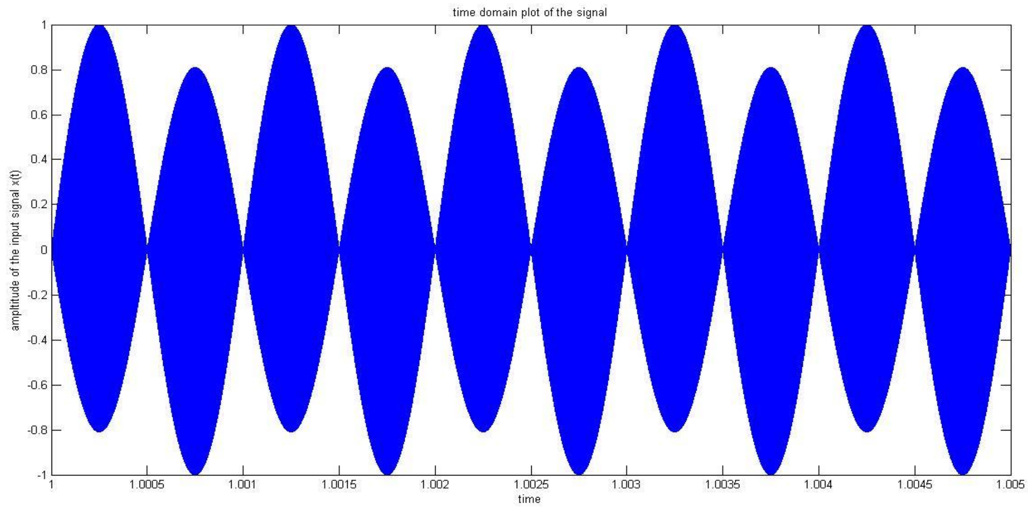


Figure 11: Time domain plot of the signal

<u>Interference Generation:</u> The interference is generated by passing $x(t)$ through a pre-distorted channel $h(t)$ which has the following properties or assumptions:

- Maximum delay that a signal undergoes is 100 units of the sampling time
- Only 16 out of these units have non-zero amplitudes and rest are all zero
- A magnitude to the non-zero components is assigned randomly and the magnitude of each is restricted to a maximum of 0.2

This pre distorted channel $h(t)$ is convolved with $x(t)$ to form the multipath $z(t)$.

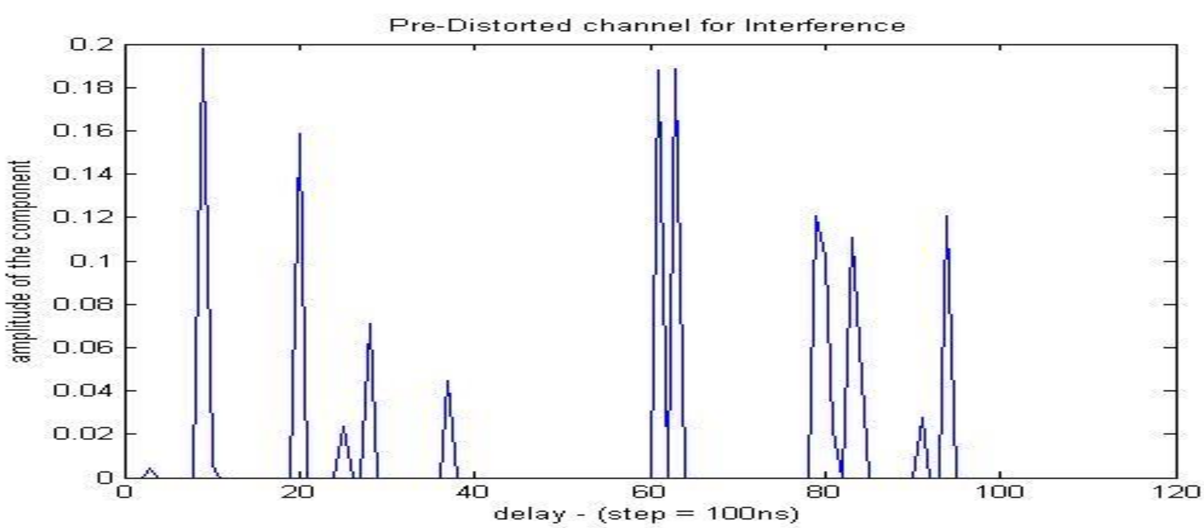


Figure 12: Pre Distorted Channel for Interference

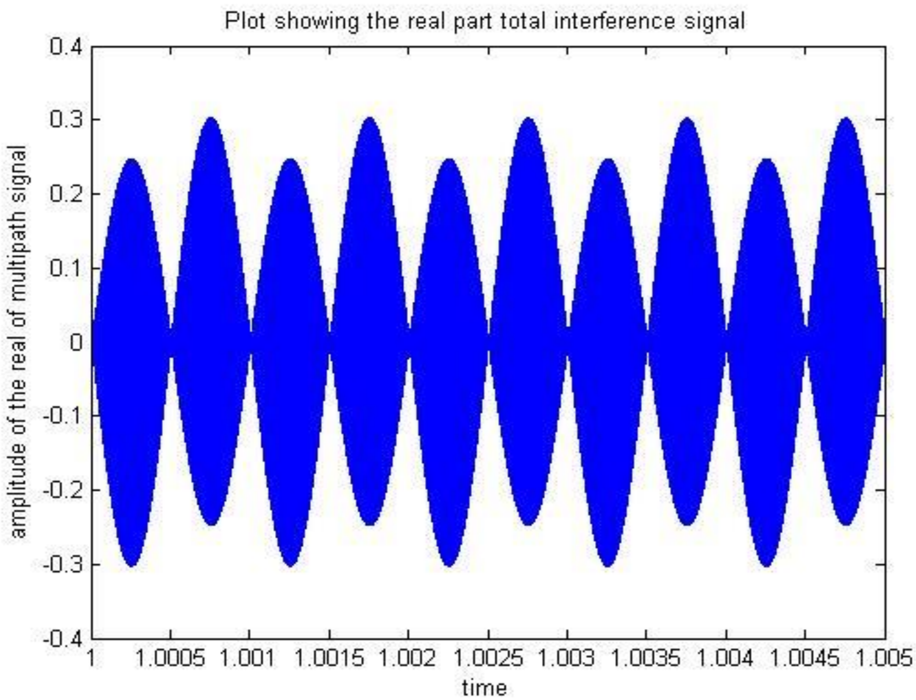


Figure 13: Plot showing the real part of the total interference signal

A particular case of $h(t)$ and the corresponding $z(t)$ is shown above

<u>Filter Response:</u> The method mentioned in the section of padẻ approximation is followed. A two stage architecture that is used is shown below:

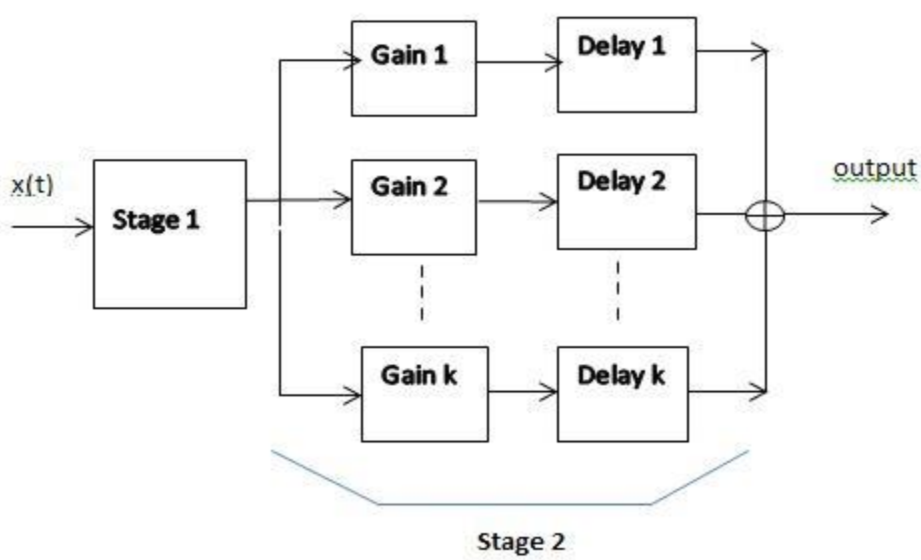


Figure 14: Tunable filter structure

The number of delay lines are chosen initially and then, value of each delay is tuuned so as to find the find the best output $y(t)$ that matches with the interference signal $z(t)$.

An example for the gain and delay method is given below.

The filter response

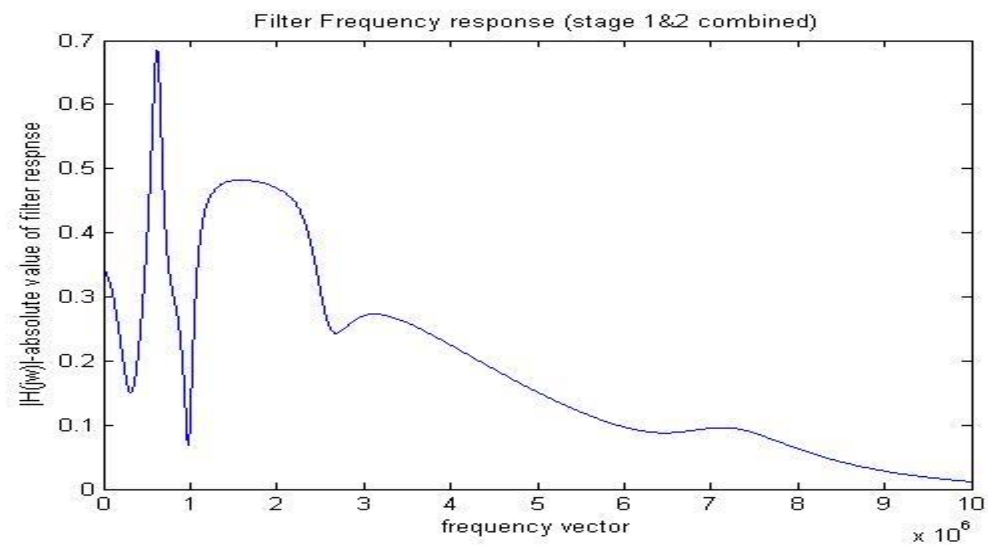


Figure 15: Total Filter Frequency response

Final output $y(t)$

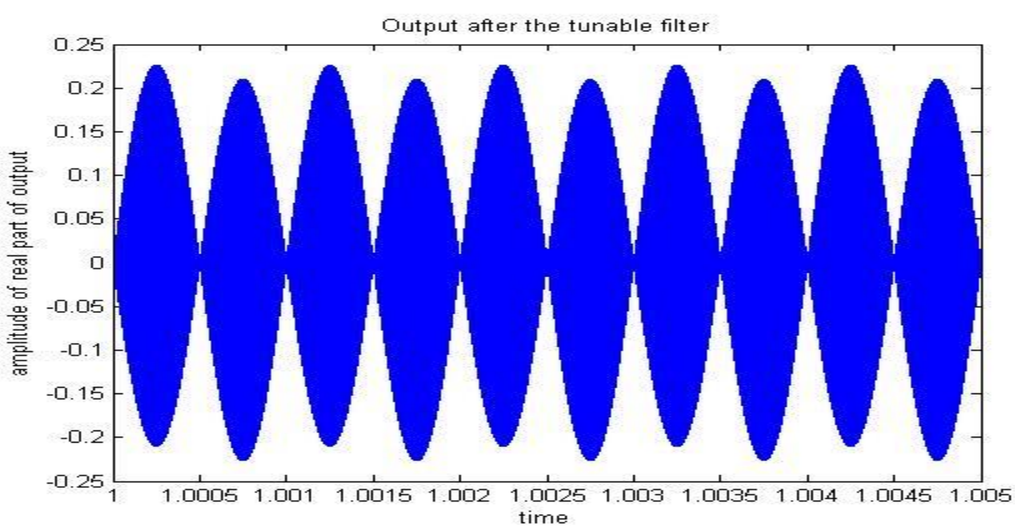


Figure 16: Output of the tunable filter

The delay is incorporated by truncating e^{-st_d} to an n^{th} order polynomial using the Taylor series approximation. So, in the case of tuning there are two sets of parameters to be taken care of : delay and truncation length (order n).

Observations from this approach:

As mentioned, a 1 kHz signal modulated on a 2 MHz carrier is used as the input and 16 taps channel is used for generating the interference signal.

Coming to the tunable filter,

The stage 1 is just to remove the noise which is outside the band of the signal and not to affect the signal ideally. So, a filter which is maximally flat over the signal bandwidth should be used. To achieve this, a 12^{th} order Bessel filter with a cutoff of 8MHz is used. Since, signal bandwidth is 2 MHz; a cutoff of 8 MHz ensures that the filter response is flat over the signal bandwidth.

In stage 2, the number of delay lines that are to be considered is to be fixed. Then, the truncation length is to be fixed. Then, the values of gains and delays are varied and the total cancellation is observed at each case.

If matched perfectly, this method gives really good cancellation (about 70 dB cancellation is observed). But, there are some drawbacks with this method. They are:

- The very first drawback is the number of variables that are to be tuned. One has no idea on how many delay lines are to be taken, what truncation length is to be used and the gain and delay values. This leads to a very large sample space to look out for a solution.
- Second reason is that the probability of ending up in a local minima is high and this might not give the exact cancellation that can be achieved
- The practical implementation of this approach is highly infeasible

So, other ways of implementing the tunable filter structure are to be looked at. This issue is addressed in the next approach where a single rational polynomial is used as a filter to cancel the interference signal.

Improvement 1:

Signal Generation and Interference Generation sections in this approach remain the same as in the above approach. Only the tunable filter section is modified which is explained below:

In this approach the following transfer function is used as a tunable filter

$$H(s) = \frac{a_0 + a_1 s + a_2 s^2 + \cdots + a_n s^n}{b_0 + b_1 s + b_2 s^2 + \cdots + b_m s^m}, m \geq n \quad (7)$$

This is an n^th order tunable filter. Now, the idea is to vary the coefficients a_i and b_i and find the best filter that gives the best cancellation for a particular input signal and interference

Initially, a specific range is taken for all these coefficients $L_1 \leq a_i, b_i \leq L_2$ with a particular step $\frac{L_1 - L_2}{N}$.

<u>Method for obtaining the best cancellation:</u>

- Get the values for L_1, L_2 and N
- For these values, generate all possible filters

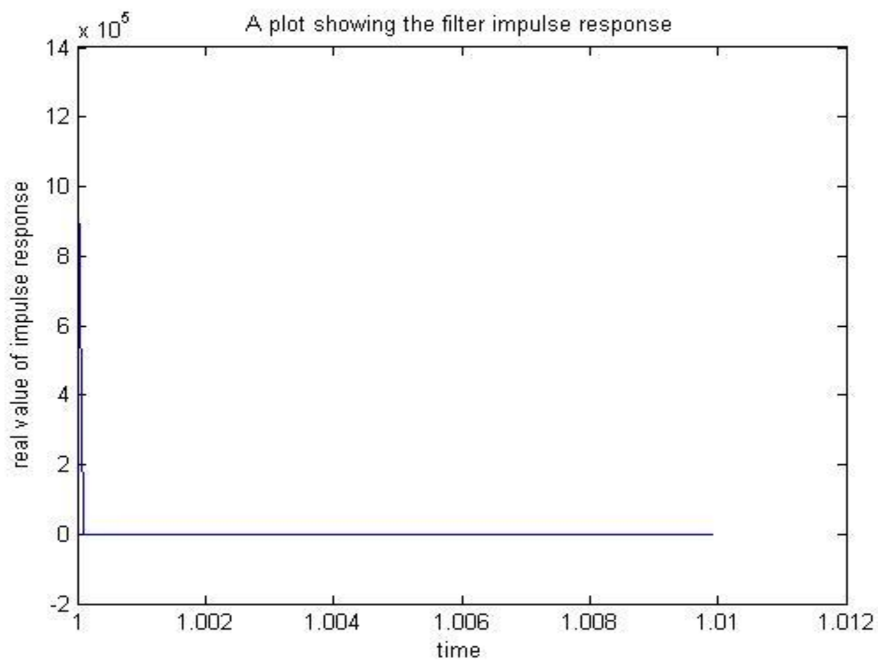


Figure 17: A figure showing impulse in time domain filter response

- Pass the input signal through all these filters, cancel the interference signal with the output of each filter and choose the one that gives the highest cancellation
- Once the highest cancelling filter is chosen, consider a very fine step in the neighborhood of that filter to check for even higher cancellation

Observations from this approach:

1. Though this approach seems better than the previous approach in terms of the number of variables, the sample space of each variable is now very large.

2. The cancellation that is obtained is very low at some times because it doesn't exist in the sample space we look for.

2. The figure 16 shows the time domain response of filter to a certain combination. Here one can clearly observe that the response resembles an impulse. This happens because the coefficients chosen are not scaled properly.

3. One more important observation is that the multipath components are given random values for a given set of delays. But, a model for the multipath should be incorporated based on the delays each component traversed

The cancellation graph for a particular case is as below:

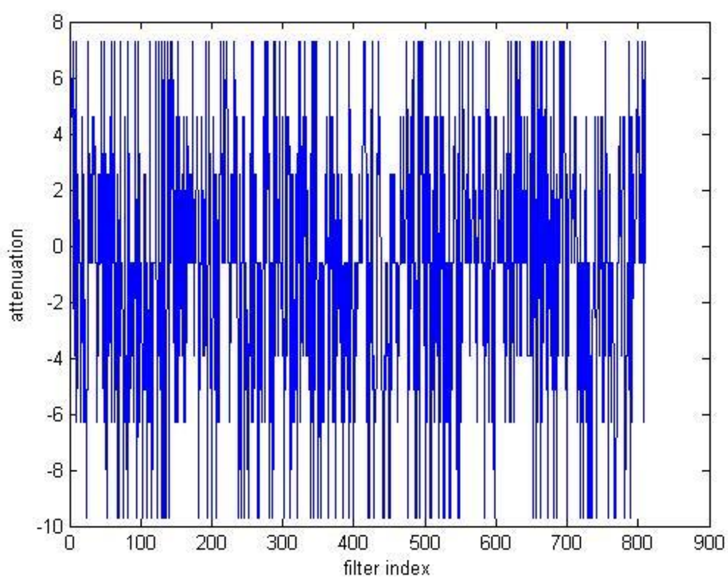


Figure 18: Cancellation for different filters

The issues in the above case are addressed in the next approaches. Though there many other approaches tried by addressing each mistake, they are not significant enough to discuss. The next approach which accounts for all the issues and still gives good results is mentioned below.

Improvement 2:

Signal Generation: $x(t)$ is a two tone sinusoid as before but using a scaled down version of frequencies is not used in this approach. All the signals have a bandwidth of few MHz and are modulated on a 2.5 GHz carrier.

Interference Generation: In the earlier approaches, 16 taps are used consistently for channel and the amplitudes are assigned randomly. This situation is highly impractical. So, now a path loss model is used which assigns a value for the amplitude based on the delay of the signal.

The model used is based on these delay and amplitude vectors:

Delay (ns)	Amplitude
0.33	0.0794
0.53	0.0501
0.84	0.0316
1.32	0.0178
2.1	0.0056

Consider the multipath components at these particular delays and amplitudes being a complex Gaussian random value with zero mean and variance as the corresponding amplitude vector value from the above table.

An example for the multipath taps with the above way of implementation is shown in the figure below.

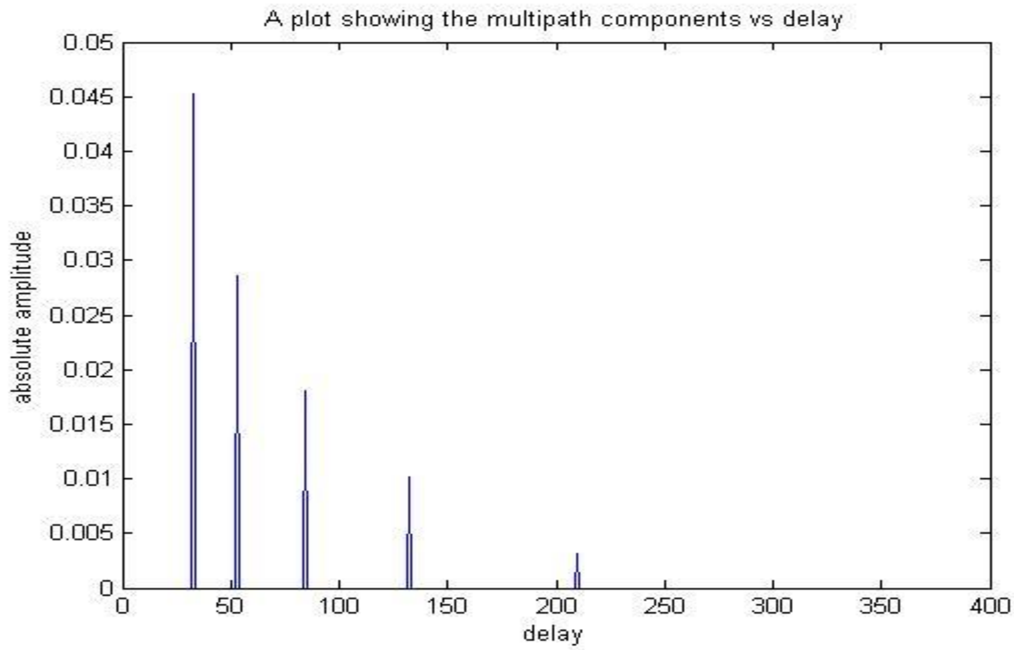


Figure 19: A plot showing multipath components vs delay

Filter response:

The basic structure of the filter is similar to that of the Approach 2.

1. One major difference is that all the coefficients of the filter are to be scaled by a factor of $2\pi f_c$.

$$a_{i,new} = \left(\frac{1}{2\pi f_c}\right)^i a_i \qquad (8)$$

$$b_{i,new} = \left(\frac{1}{2\pi f_c}\right)^i b_i \qquad (9)$$

2. The next difference is that the denominator coefficients are kept fixed. The Bessel coefficients of the same order are used as the $b_i's$.

With these differences, we are only left with 'n+1' variables to tune where 'n' is the order of the filter. An additional gain term is used for tuning in cases where a significant amount of phase matching is achieved.

Now, we look in detail the different cases and conditions under which this model is tested to analyze the cancellation that can be achieved.

Instead of going into particular examples, different categories are grouped together and these groups are looked at separately. These can be broadly classified based on the filter order.

<u>Results for 0^{th} order filter:</u>

A 0^{th} order filter simply just alters the gain of the signal. So, when 0^{th} order filter is used to cancel an interference signal, a very good cancellation can't be seen.

An example where the interference signal and the filtered signal are plotted on the same graph is shown below

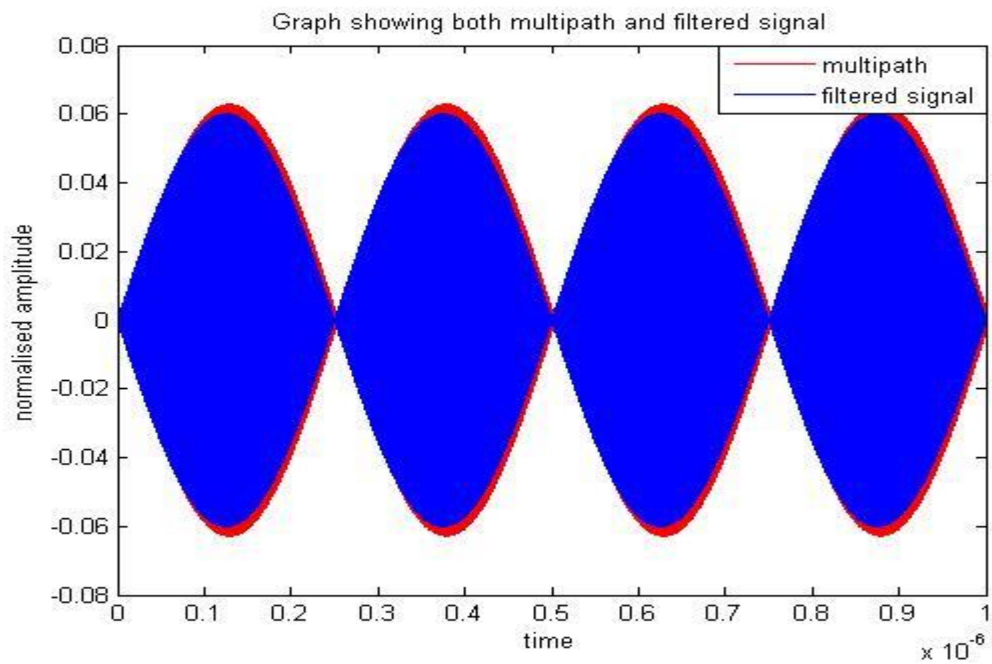


Figure 20: Graph showing both multipath and filtered signal for 0 order filter

The cancellation in this case is observed to be 22.61 dB

0^{th} order filter doesn't account for the cancellation of the derivative terms of the signal. The range of attenuations achieved in different cases range from 15 to 28 db. Later first order case is analyzed where one can see the cancellation of both signal and the first derivative term.

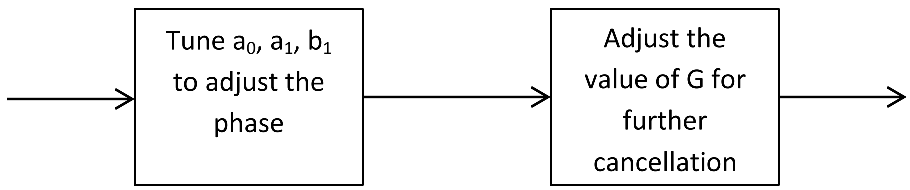
<u>Results for a 1^{st} order filter:</u>

The filter structure used for tuning in case of the 1^{st} order filter is the following

$$H_{order,1}(s) = G\,\frac{a_0 + a_1 s}{1 + b_1 s} \qquad (10)$$

Where G, a_0, a_1, b_1 are complex numbers and scaled accordingly using the mentioned equation y

Here, initially the value of G is kept low. Then, other coefficients are tuned in order to adjust the phase to the best possible extent. After this, the value of G is adjusted to further cancel the signal.



The cancellation achieved is actually a function of the amplitude of the multipath. So, limit on the maximum interference is considered for simulation. Here, the maximum absolute value of the multipath is limited to be 0.2 times the maximum amplitude of the input signal.

From now, whenever the term amplitude of the multipath is used, it is assumed to be normalized with respect to the input signal.

The table below mentions the cancellation achieved as a function of the amplitude of the multipath for some cases. (Number of multipath taps is 5 here)

Amplitude of the multipath	Cancellation achieved (in dB)
0.016	37.52
0.033	33.62
0.071	32.7
0.0832	31.3
0.098	28.16

Results for 2nd order filter:

Second order filter should account for even higher cancellation because it can cancel even second derivative terms that are present.

The filter structure is given below:

$$H_{order,2}(s) = G\,\frac{a_0 + a_1 s + a_2 s^2}{2+\sqrt{3}s+s^2} \qquad (11)$$

The denominator is a second order Bessel filter with cutoff 1 Hz. So, these coefficients have to be scaled according to the cutoff required.

So, even in this case, there are only 4 variables.

This results for cancellation achieved for 2nd order filter are mentioned below

Amplitude of the multipath	Cancellation achieved (in dB)
0.016	44.63
0.033	39.75
0.071	37.38
0.0832	34.15
0.098	32.37

All these cancellations are plotted in the same graph as below

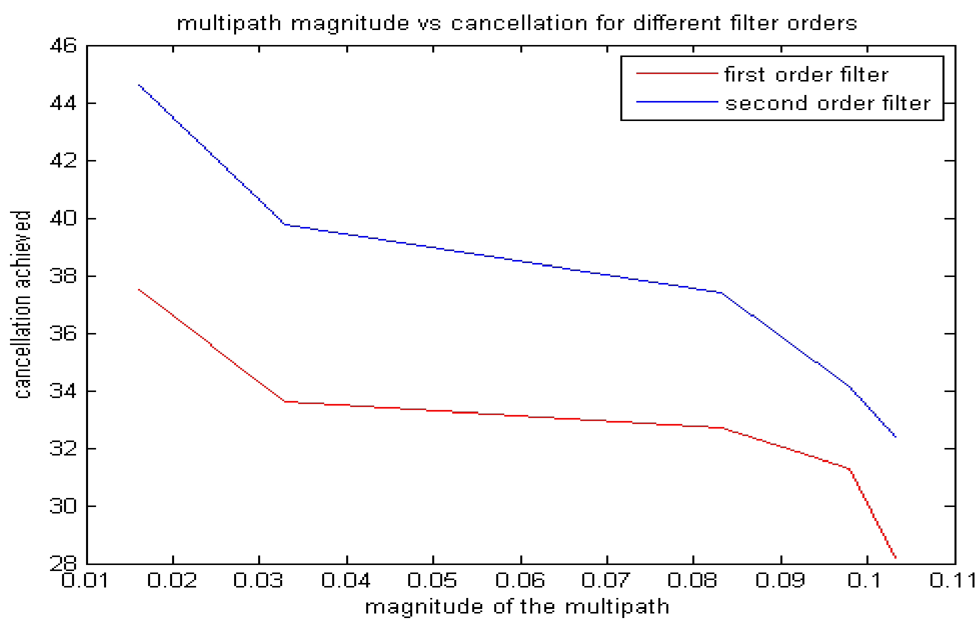


Figure 21: Cancellation achieved for first and second order filters

Improvement 3:

In the above approach, for the case of 2nd order filter, there are only 4 parameters that can be tuned. The denominator is kept fixed as a Bessel polynomial.

In some cases where the cancellation is less, the variability of the denominator coefficients has showed improvements. It was finally understood that the variability in the denominator helps for fine better cancellation.

Even from the implementation aspects, building a cascade of n first order filters is actually easier than the direct nth filter.

On the basis of these reasons, the filter structure is modified as to below:

$$H(s) = G.\frac{\prod_{k=1}^{n}(s+z_k)}{\prod_{k=1}^{n}(s+p_k)} \qquad (12)$$

So, an nth order filter has $2n + 1$ tuning parameters.

The results for sinusoid waveforms using this filter structure are given below for a second order system.

Amplitude of the multipath	Cancellation achieved (in dB)
0.016	55.26
0.033	43.48
0.071	41.3
0.0832	39.11
0.098	35.19

This approach is tried out for even higher order filters, but there is a significant cancellation improvement only till the 3rd order filter case. The cancellation achieved for different filter orders at different multipath amplitudes is shown in the figure 21

All the above results are achieved by using continuous tuning. But, when these filters are implemented practically, only discrete values for the coefficients can be used. So, a graph showing difference between continuous and discrete tuning is shown in the figure 22

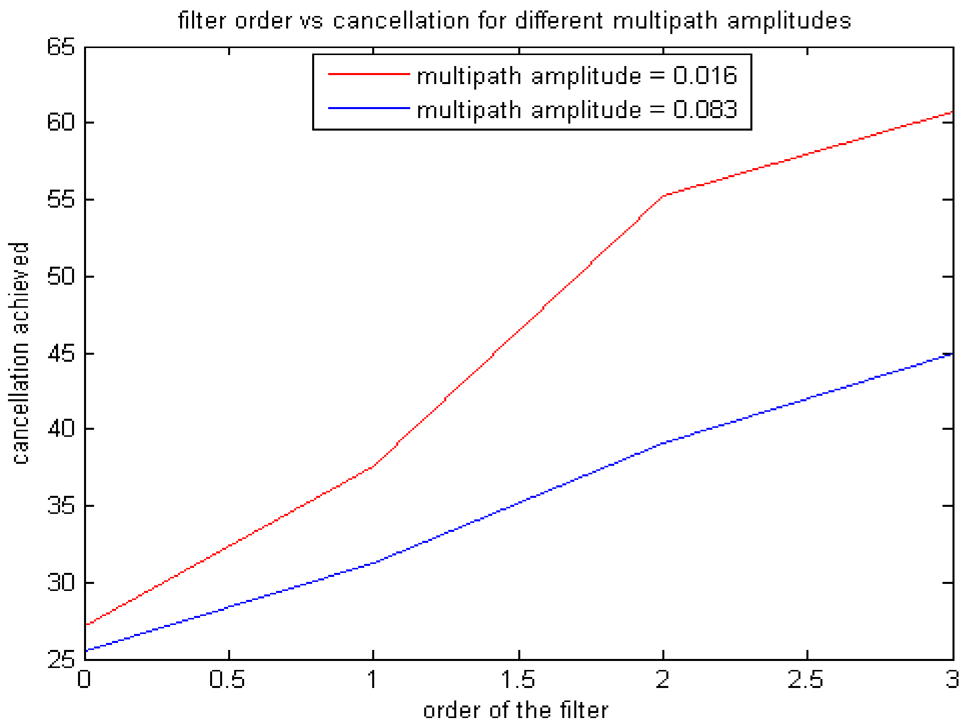


Figure 22: filter order vs cancellation at different multipath amplitudes

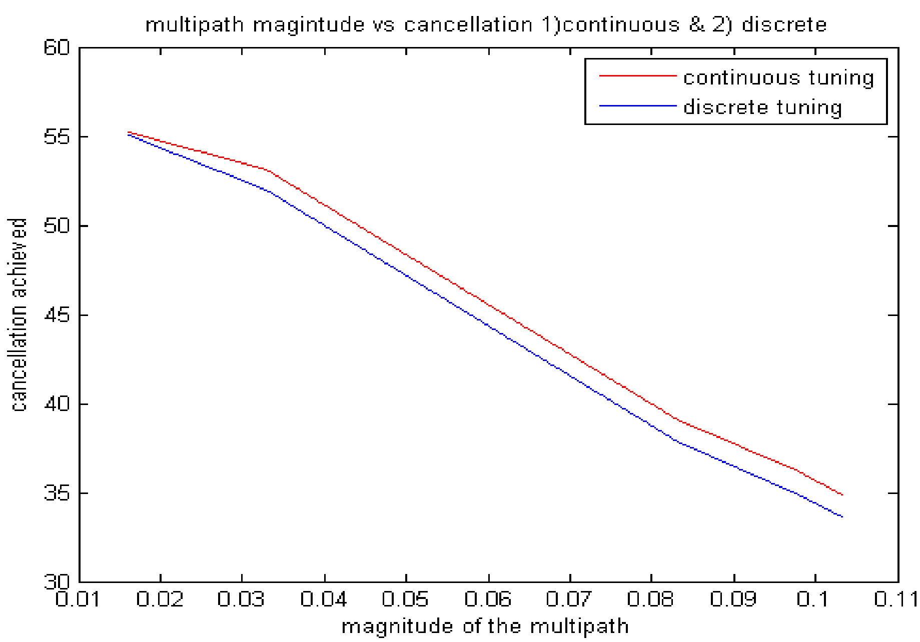


Figure 23: Multipath magnitude vs cancellation for continuous and discrete tuning

All the analysis reported till now is for a two tone sinusoidal signal input. The above method is extended to OFDM signals which are the actual signals that will be finally used for transmission and reception.

Results for OFDM signals:

Signal Generation: OFDM samples are generated as per the method mentioned in theory section.

Having generated the OFDM digital samples, they can be scaled to any bandwidth. After adjusting the signal to a particular bandwidth, it is modulated on a 2.5 GHz carrier.

The OFDM samples used are as shown in the figure below:

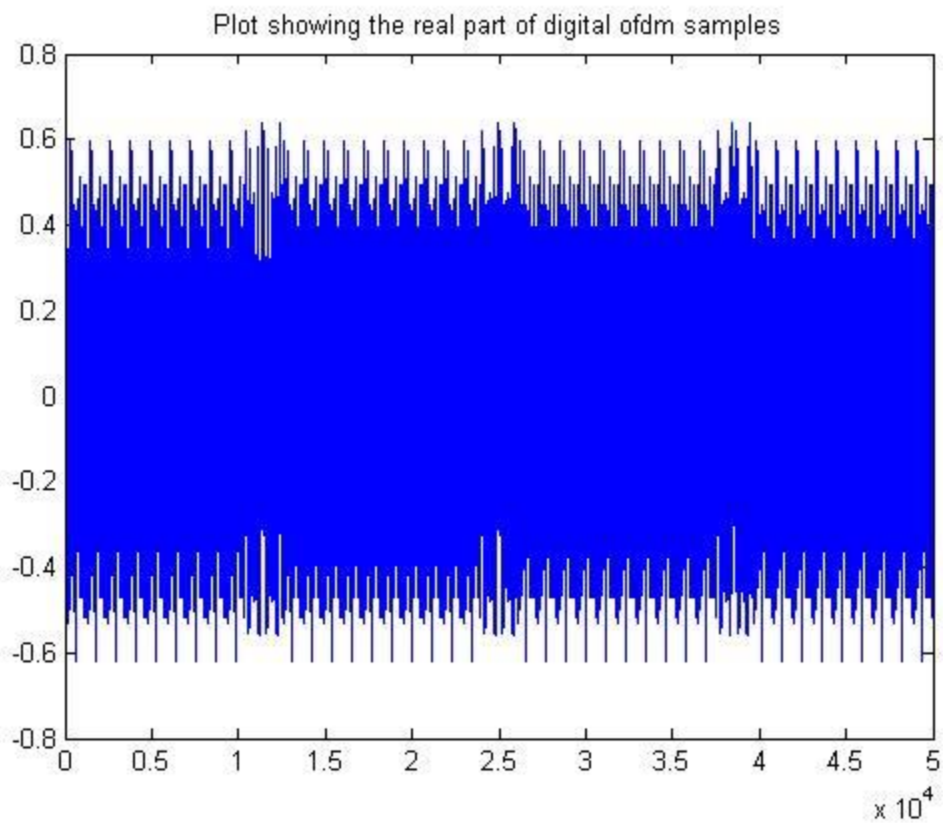


Figure 24: Plot showing the real part of digital OFDM samples

One of the issues addressed during this process is mentioned below:

- To assign a bandwidth to these digital samples time division between the samples must be the inverse of the required bandwidth.
- But, for those values of time divisions the value of the carrier multiplying them is always 1. So, actually the modulation is not happening.
- If the time division is changed to have more carrier samples, it even changes the bandwidth of the signal.

So, the solution is to first assign the time division to the samples as the inverse of the bandwidth. Then, interpolate the signal by a factor L which allows for the proper modulation to happen. Then, modulate the resultant signal with the carrier.

<u>Example:</u>

Say, a 10 MHz OFDM signal is to be modulated on a 2.5 GHz carrier. Sampling rates for signal & carrier would be 10 MHz and 2.5 GHz respectively. So, the minimum sampling time is 100 ns for the signal & 0.4 ns for the carrier.

The carrier is oversampled by 2 and so, a sample at every 0.2 ns is needed.

So, in this case interpolation factor, $L = \frac{sampling\ time\ of\ signal}{sampling\ time\ of\ carrier} = \frac{100\ ns}{0.2\ ns} = 500$

This method is to be followed in general for any other OFDM signal modulation.

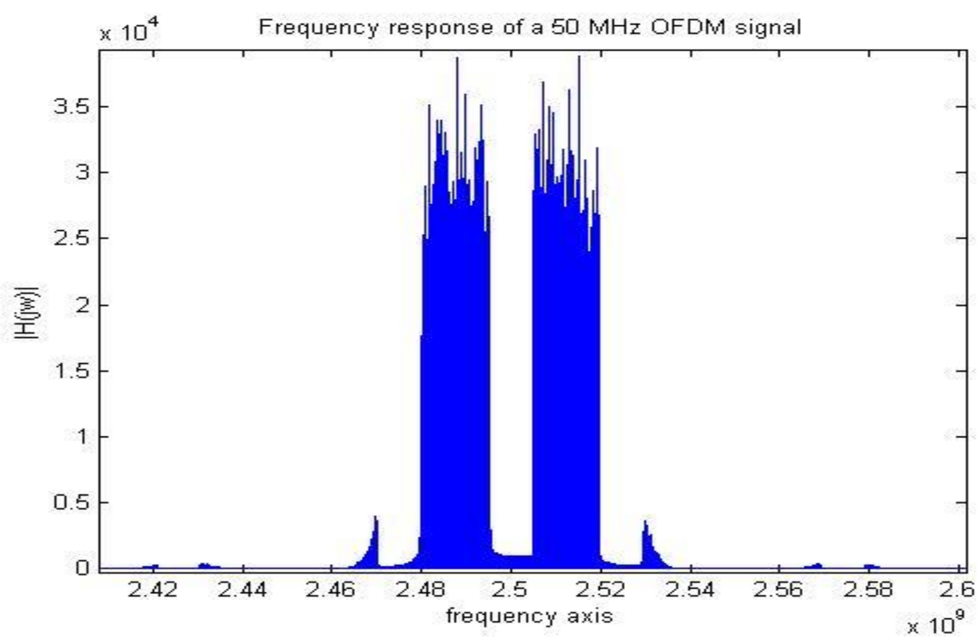


Figure 25: Frequency response of a 50 MHz OFDM signal

Frequency response of a carrier modulated OFDM signal is shown in the figure 24.

The method followed for the interference signal generation and the tunable filter operation is same as mentioned in the previous section. But, in case of tunable filter, only discrete tuning of coefficients is used.

<u>Details on discrete tuning:</u>

- 64 steps of 0.5 dB step size
- Covering a range of -22 dB to 10 dB

A particular example for the case of cancellation using OFDM signals:

- The signal used is a 10 MHz OFDM signal.
- Number of taps for multipath = 5
- The maximum normalized amplitude of interference signal = 0.095
- A second order tunable filter is used. The coefficients obtained for the highest cancellation are poles = -4 dB , -11 dB ; Zeros = -5.5 dB , -1.5 dB
- The cancellation achieved for this case is 42.31 dB

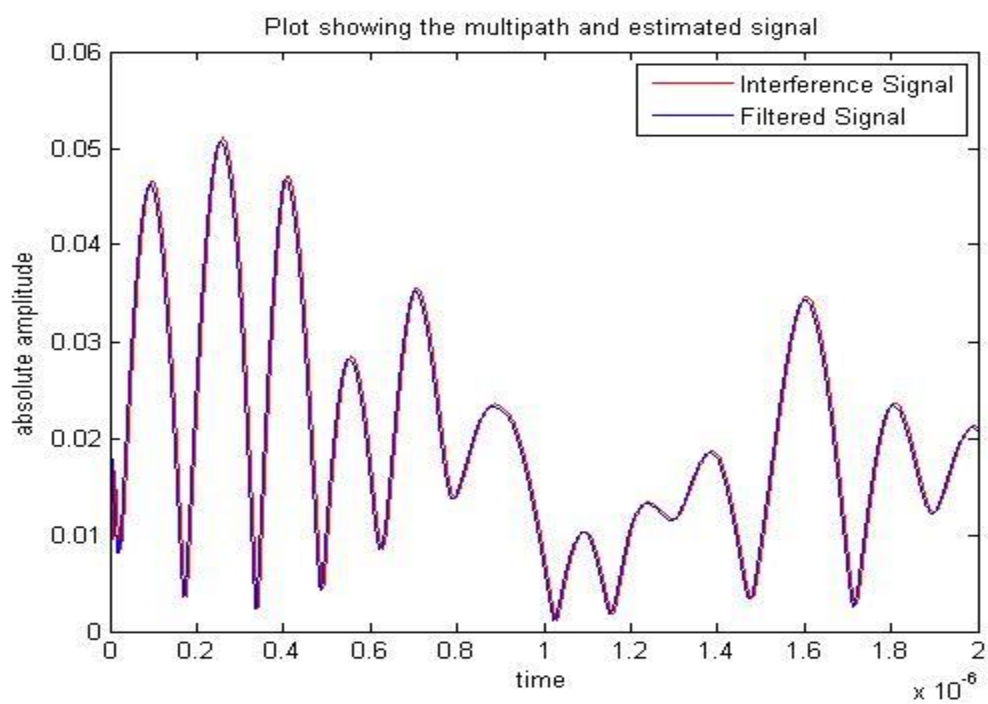


Figure 26: Plot showing the multipath and filtered signal

This graph shows a part of interference and filtered signals in time domain.

Results obtained at different signal and interference conditions are reported below.

1. Order of the tunable filter vs Cancellation achieved:

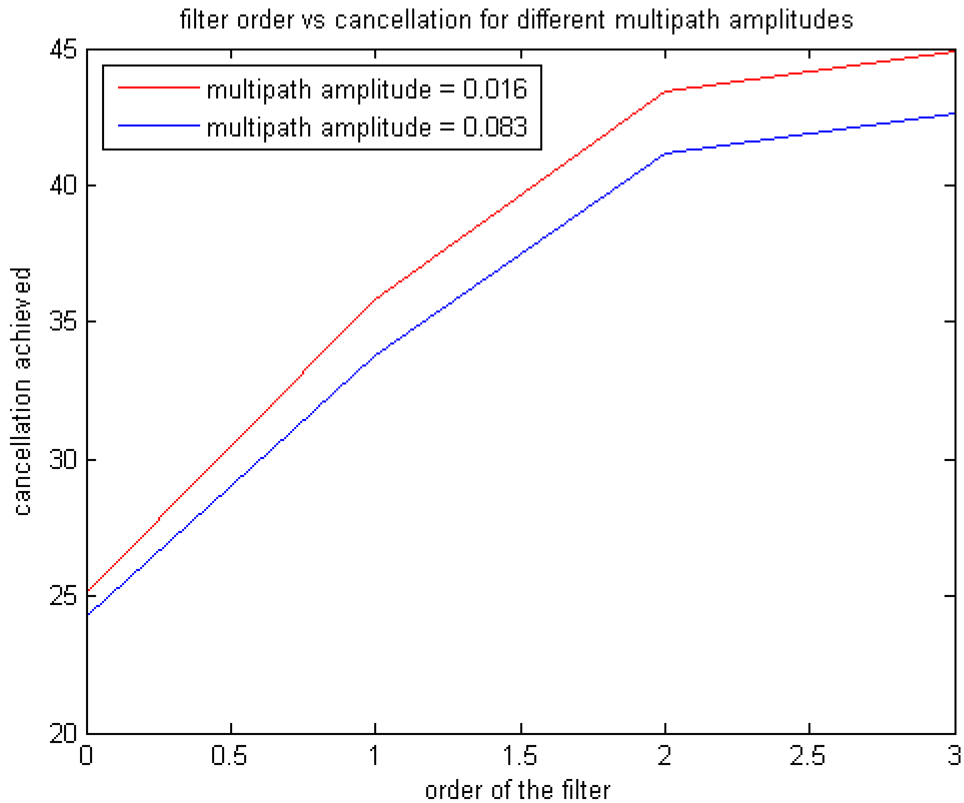


Figure 27: filter order vs cancellation at different multipaths

The cancellation for a particular interference signal is calculated for 0^{th}, 1^{st}, 2^{nd} and 3^{rd} order filters.

This is done at 10 MHz OFDM signal and the number of paths used for interference generation is 4.

Observation:

As the filter order is kept on increasing, there is an improvement in the cancellation achieved.

But, this increase is not linear. One can clearly see a saturation appearing with the increase in the filter order.

This is consistent with the initial assumption that the contribution of the higher order terms in the interference signal is very low.

2. Number of Interference paths vs cancellation achieved:

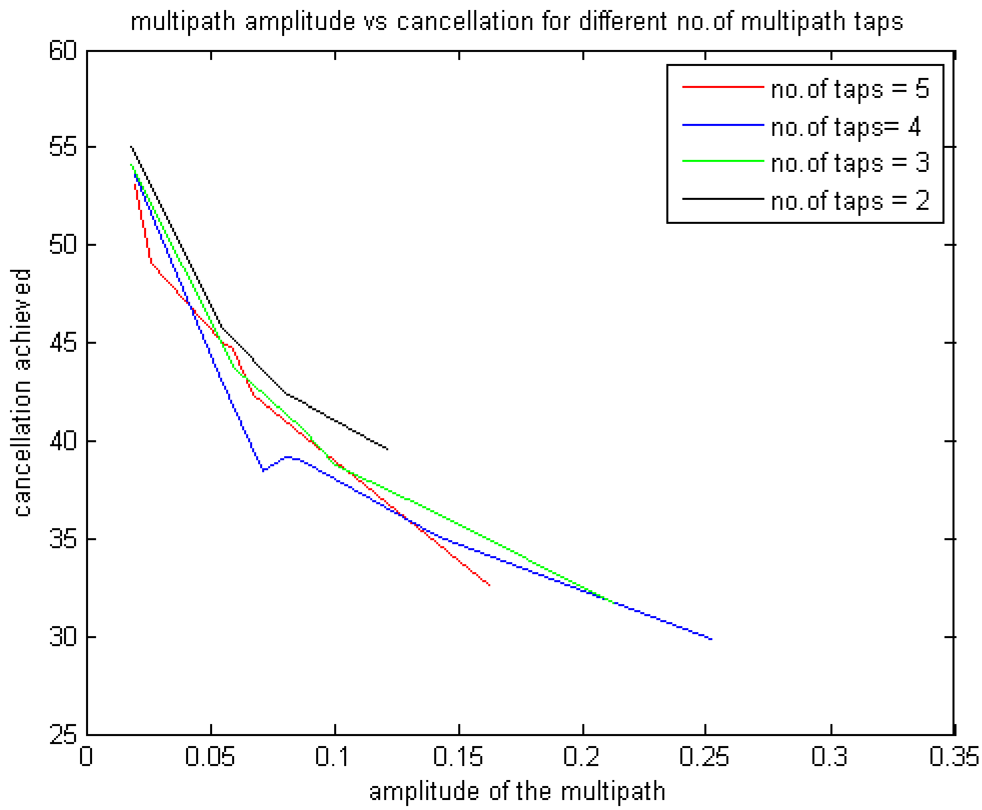


Figure 28: Multipath amplitude vs cancellation at different no.of multipath taps

This analysis is done for a 10 MHz signal and a 2nd order tunable filter is used.

This graph shows that the cancellation achieved varies with the number of interference paths but,

- Variation is very small when the normalized amplitudes of the interference signal are low
- A significant variation is observed only at higher amplitudes of Interference
- Trend observed is that lesser number of paths gives better cancellation

3. Amplitude of interference vs cancellation achieved:

This analysis is done for a 10 MHz OFDM signal a 2nd order tunable filter.

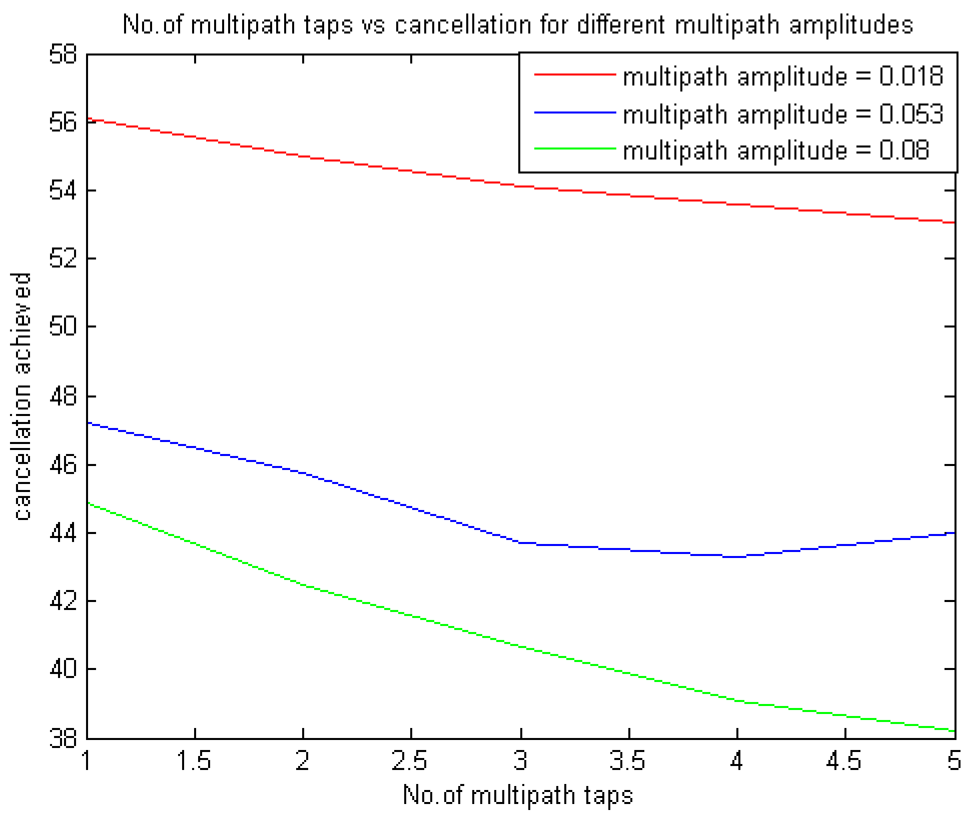


Figure 29: No.of multipath taps vs cancellation at different multipath amplitudes

Observation:

Lower the amplitude of the interference, higher is the cancellation that is achieved.

Also the observation made in the previous graph is visible here. The slope of the variation is less steep for the case with less number of interference paths. So, this makes sure that the following analysis is consistent with before mentioned results.

4. Bandwidth of the signal vs Cancellation achieved:

The analysis is done under the following conditions:

- Number of interference paths used = 4
- Order of the tunable filter used = 2nd order
- The cancellation is calculated for 10 MHz , 20 MHz , 25 MHz, 40 MHz and 50 MHz signals

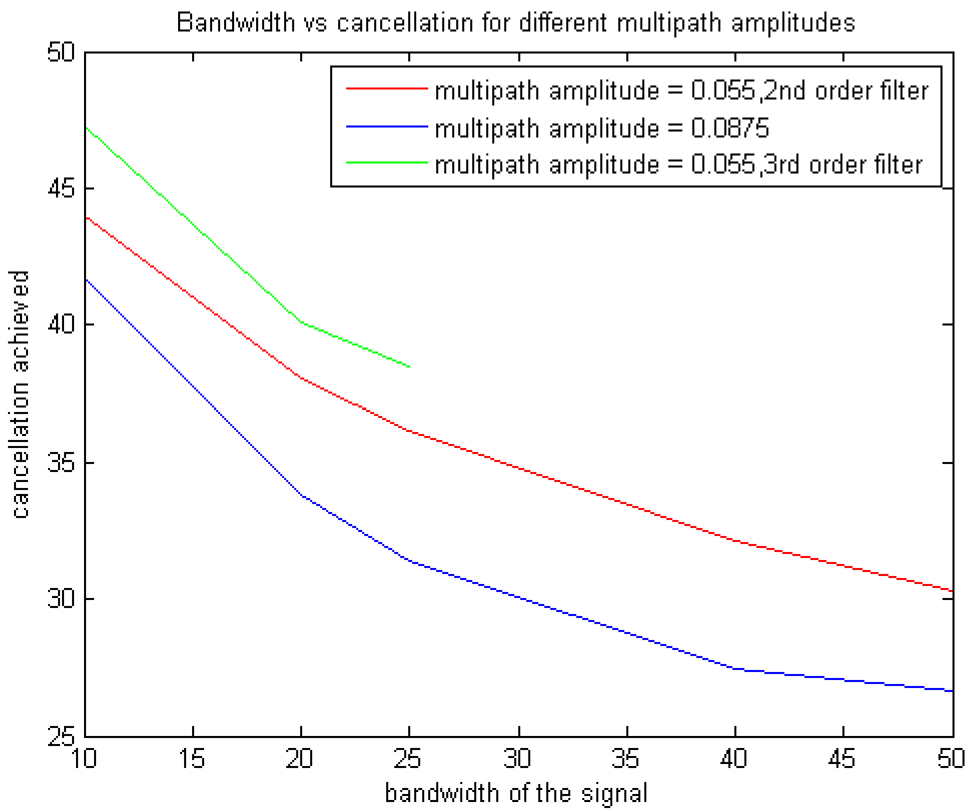


Figure 30: Bandwidth of signal vs. cancellation at different multipath amplitudes

Observation:

The cancellation achieved decreases significantly with the increase in the bandwidth.

This says that more number of higher order terms should be considered to get better cancellation for the signals having higher bandwidths.

The shape of the graph looks similar at the two different interference amplitudes which says that the cancellation variation with the bandwidth of signal is an independent variation.

The results obtained for cancellation when compared with that of Taylor approximation based results are observed to the better. A comparison of the measured results in case of Taylor and simulation results in case of the Padė are analyzed below.

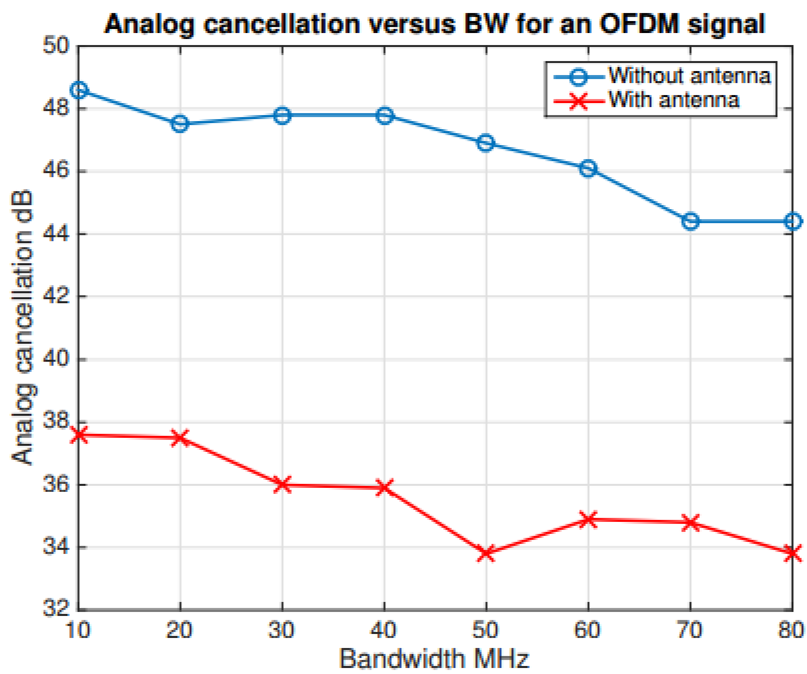


Figure 31 Cancellation vs. Bandwidth for the case of Taylor Approximation

These are the results for the cancellation using Taylor Approximation. These results include the cancellation provided by the isolator. Coming to the results of cancellation using Padė Approximation, from the Figure 30, these are the observations:

For a 10 MHz OFDM signal,

Taylor Approach: 49 dB of cancellation

Padė Approach with 2nd order filter: 41 dB + 18 dB = 59 dB

Padė Approach with 3rd order filter: 45 dB + 18 dB = 63 dB

For a 20 MHz OFDM signal,

Taylor Approach: 47 dB of cancellation

Padė Approach with 2nd order filter: 37 dB + 18 dB = 55 dB

Padė Approach with 3rd order filter: 40dB + 18 dB = 58dB

So, there is a definite improvement by shifting to Padė approach from the Taylor. This ends the discussion on the simulation results of the tunable filters for different signal, interference and filter conditions.

5. Practical Implementation of tunable filters

The filter that is to be built is the following form:

$$H(s) = G.\frac{\prod_{k=1}^{n}(s+z_k)}{\prod_{k=1}^{n}(s+p_k)} \qquad (13)$$

Consider a simple case of a first order filter:

$$H(s) = G.\frac{s+z_1}{s+p_1} = \frac{Gs+Gz_1}{s+p_1} = \frac{k_0 s+k_1}{s+p_1} \qquad (14)$$

This has 3 parameters that can be varied for the required frequency response. The simplest structure which strikes when thought of a first order filter is shown below:

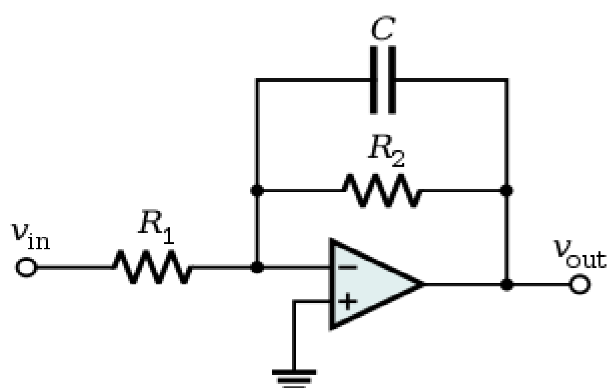


Figure 32: A first order opamp based integrator

The transfer function for the above circuit is

$$\frac{v_{out}}{v_{in}} = -\frac{R_2}{R_1}\left(\frac{1}{1+sCR_2}\right) \qquad (15)$$

But, a ks term should be introduced into the numerator. The final transfer function should follow the following flow graph:

So, according to the below flow graph, a ks term and variability of the coefficients are to be added to the figure q in order to make it a first order tunable filter.

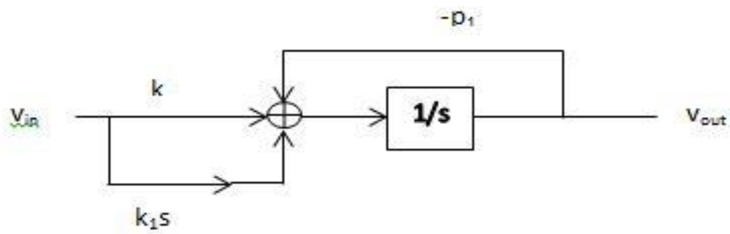


Figure 33: A signal flow graph for a first order filter

Gm-C filters:

These are class of filters where one can use variable Gm as a tuning option. These filters are advantageous for this project because of their characteristics that are mentioned below:

- Faster than active RC filters since they use only open loop stages
- Can operate at very high frequencies ranging to tens of GHz
- Low power because the active blocks drive only capacitive loads
- Linear relation between the input voltage and output current
- Processing of multiple inputs is very easy

The circuit of a first order tunable filter using Gm-C approach is shown in figure r.

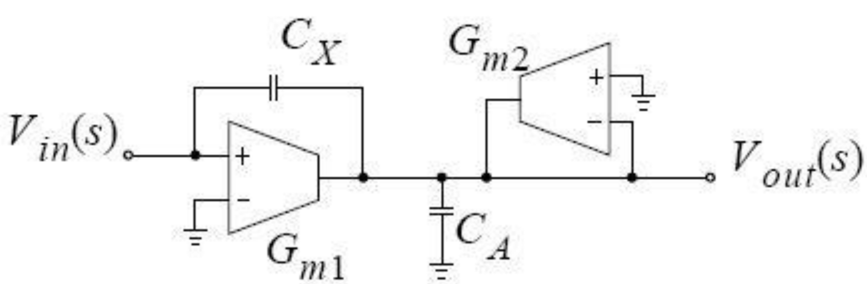


Figure 34: A circuit showing first order Gm-C filter

The transfer function for the above circuit is:

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{sC_X + G_{m1}}{s(C_A + C_X) + G_{m2}} \qquad (16)$$

By comparing the above equation with the filter equation mentioned initially,

$$k_0 = \frac{C_X}{C_A+C_X}; k_1 = \frac{G_{m1}}{C_A+C_X}; \ p_1 = \frac{G_{m2}}{C_A+C_X} \qquad (17)$$

There are 3 parameters and 4 variables. So, a relation can be assumed between any of the 4 variables C_A, C_X, G_{m1}, G_{m2}. To have an independent tuning option for all the 3 parameters, a better relation would be to have $G_{m1} = K.G_{m2}$, K is a constant.

If an n[th] order Gm-C filter is to be designed, a cascade of n first order filters can be used. So, in order to build the above circuit, tunable transconductors and capacitors are required.

A 2[nd] order Gm-C filter designed according to the above method looks similar to the figure shown below:

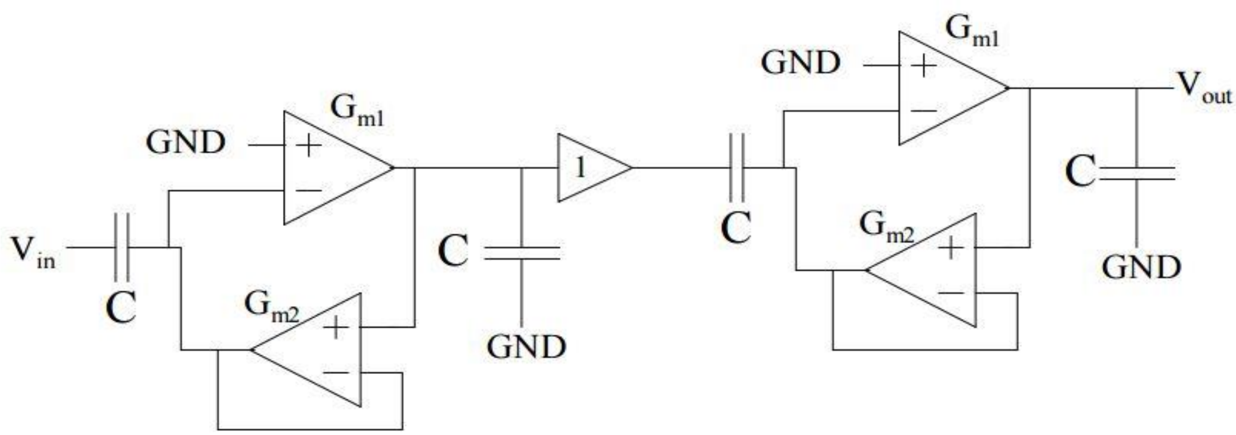


Figure 35: A circuit showing cascaded Gm-C filter

A component study has been done on tunable transconductors and tunable filters from various company products in order to find out the required components to build these circuits practically. A summary of that study is mentioned below.

Component Study:

The parameters that are considered when looking for components are:

- Maximum bandwidth of operation
- Range of gain values that can be achieved
- Value of the minimum gain step

The options of components that are found from the study are tunable transconductors , tunable VGA filters, and digital control VGAs. But, the first two components do not support the required bandwidth of operation. A few component numbers which are studied are mentioned below highlighting each on the parameters that are of interest.

Component name	Type of component	Min Freq(Hz)	Max Freq(Hz)	Step gain(dB)	Min. gain(dB)	Max. gain(dB)
ADRF6520	Baseband programmable VGA filter	0	1.25G	-	-7	53
HMC625B	Digital control VGA	0	5G	0.5	-13	18
HMC742A	Digital control VGA	500M	4G	0.5	-19.5	12
ADL5243	Digital control VGA	100M	4G	0.5	-1.2	31.3
HMC425A	Attenuator	2.2G	8G	0.5	0.5	31.5

From the above components, it is understood that attenuators cannot be used for the present purpose. So, the options left over are VGA filters and VGAs. If VGAs are used to design the filter, the best among mentioned ones would be HMC742A. Because it is very close to the specifications considered for the analysis pertaining to this project.

6. Conclusion & Future Work:

This project has addressed the possibility for Self Interference cancellation in the RF domain using Padė Approximation. All the relevant material was thoroughly studied and the simulations were performed to show the cancellation that can be achieved under different conditions.

A broad view of variation of cancellation with parameters like filter order, interference magnitude, number of interference paths and bandwidth of the signal is presented based on the detailed simulations

A part of work is done on the practical implementation of these filters but, it is limited to only identifying the circuit that can be used and looking for the components that can be used to build the circuit. So, this forms the future work.

One should take up from the last table listed, see the feasibility of building the circuit with those components and if so, building the circuit and test the circuit for all the characteristics mentioned in the simulations. With this, the discussion of this project is concluded.

7. References

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