Feedforward Interference Cancellation System
Applied to the
800 MHz CDMA Cellular Band

By

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of the requirements for the degree of

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Alain Roussel B. Sc. E.

In Partial fulfillment of the requirements
for the degree of Master of Applied Science

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Abstract

A feedforward noise cancellation system (FFCS) prototype was developed to complement a cellular base station duplexer of reduced order and reduced size operating in the CDMA 800 MHz cellular band. The FFCS is equipped with an adaptive control system that is trained on a pilot tone signal. The pilot tone signal is used to control the center frequency of the cancellation notch of the FFCS. A cancellation performance of 50 dB was achieved in a bandwidth of 4.5 MHz with the prototype architecture. The FFCS has potential applications in communication systems requiring tunable notch filters with wideband performance using a low complexity architecture. The technology has potential for implementing a frequency agile duplexer.

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Chapter 1 – Literature Review

1.0 – Introduction

Wireless networks are easy to deploy for service in remote and hard to reach areas. Internet, multimedia and data services demand has gone up mainly influenced by the explosive growth of the Internet. New multimedia services require higher bandwidth and more advanced protocols, offering better quality of service and information security. However, the proliferation of cellular and personal communication networks burdens an already overcrowded frequency spectrum.

State-of-the-art radio access technologies (CDMA, WCDMA), provide an increase in the number of users that can be supported per unit bandwidth compared to earlier technologies of AMPS and TDMA, and requires a generous amount of the frequency spectrum. To meet the demand for new services and use the spectrum as efficiently as possible, the frequency spectrum regulating authorities impose strict rules on signal characteristics being transmitted. Spectral masks are used to define the allowable emissions from communication transmitters in licensed frequency bands. These rules impose strict requirements on the communications industry developing base station transmitters.

Cost reduction of network equipment is a key driving force behind the development of new technologies. Reducing the size and increasing the performance of equipment also carries a benefit by reducing the material costs and easing the deployment due to reduced real-estate requirements for the equipment. Cost and size reduction coupled with an increase in performance are significant challenges in the improvement of current technologies and development of new 3G communication systems.

The cellular base stations in use today were not developed to cope with the rigors of the most modern of the developing technologies. As high capacity data technologies are

implemented in mobile networks, base station architecture need to evolve to meet the changes. Wireless solutions must remain affordable, amenable to mass production and easily deployable.

Wireless communication base stations require the collocation of a high power transmitter and a sensitive receiver. Modern full duplex base station units operate with dual band wireless communication standards, such as cdmaOne (IS-95) and cdma2000 (IS-2000). The radio front-end of the base stations use only one main antenna to receive and transmit signals. Some base stations will also use a diversity antenna. Collocation of a high power transmitter and sensitive receiver places severe demands on the duplexer. The duplexer is the first stage of the receive section following the antenna and the last stage of the transmit section preceding the antenna in the radio front-end of the base station.

Figure 1.0.1 illustrates a typical radio front-end used in current cellular base stations. The modulator modulates the baseband information signal and up-converts it to a transmit channel broadcast frequency. The modulated signal is amplified by a power amplifier (PA) and filtered by the duplexer before being transmitted over the air channel by the RF antenna. A signal received by the RF antenna is filtered by the receive channel of the duplexer before being amplified by a low-noise amplifier (LNA). The demodulator down-converts the received signal to an intermediate frequency (IF) or baseband to extract the information.

Wireless communication standards impose strict regulations and performance requirements on base station radio front-end designs. To meet these requirements filters with tight performance characteristics and highly linear power amplifiers (PA) must be used. Multi-carrier power amplifiers (MCPAs) are now used instead of combining the output of several single carrier amplifiers. MCPAs benefit from reduced power consumption (increased efficiency), reduced size and reduced power dissipation compared with combining several single carrier power amplifiers. However, the main

duplexer of current cellular base stations still contributes a significant cost and size to the base station radio front-end.

Figure 1.0.1 – Typical Cellular Base Station Radio Front-End

Modern CDMA base stations are driving the duplex filter technology in the areas of: size reduction, high performance in terms of transition band sharpness (less than 3% bandwidth of operating frequency band), low passband insertion loss of less than 1 dB, high rejection band attenuations in the range of 80-90 dB, low cost and frequency agility. To satisfy these demands, new concepts and advances in circuit design are needed to replace or enhance the main components of a communication system radio front-end. The components of a communication system that may be improved upon include the duplexer, RF amplifiers, mixers, intermediate frequency (IF) filters and local oscillators (LO).

This research, done in collaboration with Nortel Networks' Wireless Technology Laboratories, was aimed at investigating the possibility of applying a new technology in the radio front-end of current base station technologies. The new technology is targeted at size and cost reduction of the duplexer. The proposed approach makes use of an active circuit to complement the duplexer.

An adaptive feedforward noise cancellation system (FFCS) was developed to complement a duplex filter. The FFCS consists of a microwave interferometric network used in place of a band-reject filter in the main RF path of a communication channel. In the context of a duplexer application, the FFCS has the same function as a bandpass filter used to reject the receive-band noise present in the transmit path before it reaches the receiver, without disturbing the desired transmitted signal. The result is a higher signal-to-noise ratio at the input of the receiver.

The thesis is organized in 5 chapters. **Chapter 1** presents motivation for this research by introducing the current challenge posed by modern wireless communication standards on the base station radio front-end design. Arguments for choosing feedforward technology over other technology options as a solution are discussed.

Chapter 2 presents the necessary theory behind feedforward noise cancellation. The concept of cancellation performance is introduced and the main cancellation performance determining factors are analyzed. The system level architecture used for the prototype is developed in detail. The performance analysis and control system topology establishes the requirements for the hardware implementation.

In **Chapter 3**, the hardware implementation is explained in detail. Measurement results on the hardware prototype are presented in **Chapter 4**. **Chapter 5** proposes applications for the FFCS and suggests further areas of research based on the findings of the present thesis.

1.1 – Thesis: Theory Maintained Through Argument

Throughout this thesis, the author maintains that a microwave interferometric network can be used to implement a compact, state-of-the-art actively tunable RF notch filter. The microwave interferometric network can be equipped with an adaptive control system to optimize the control variables affecting the performance. The control system can track

changes in the control input within the tracking bandwidth of the control loop, such that the operating conditions for optimal performance are maintained. The control input can be used to render the filter frequency agile.

1.2 – Motivation for Feedforward Noise Cancellation as Application to Modern Data Communication Systems

In communication systems operating at radio frequency (RF) and microwave frequencies, information signals travel from one point to another in the form of electromagnetic waves. The path used by the microwave energy is called the communication channel. In the electromagnetic spectrum, signals commonly carry information in terms of amplitude, phase and frequency. In practice, there is no perfect communication channel. Any information signal passing through a communication channel will be subject to modifications and distortions due to parasitic disturbances. Furthermore, a communication channel always contains a certain amount of undesired electromagnetic energy. The undesired electromagnetic energy is generated by the transmitter, the receiver or interfering signals from other systems, and is commonly considered as “noise”. The terms “noise” and “interference” are used interchangeably throughout this document and refer to the undesired spectral components falling in the frequency band of interest. The ratio of the information carrying power (desired spectral components) to the noise power (undesired spectral components) at a given point in a given bandwidth is commonly referred to as the Signal to Noise Ratio (SNR) in communication systems.

The term “noise” as used in the context of this document refers to any undesired spectral components above the thermal noise floor which fall within the receive band of a communication receiver. The undesired components may include any source of human-generated interference, such as noise from ignition sparks, undesired pick-up of spurious signals, erratic disturbances from electrical storms in the atmosphere, or any other spectral components that do not consist of a desired information carrying signal and

ultimately contributes to reducing the signal to noise ratio (SNR) in the communication channel to which the developed technology is to be applied.

The major noise contribution considered for the focus of this research results from intermodulation distortion (IMD) products and other emissions present at the output of the non-linear multi-carrier power amplifier (PA) used in the transmitter of a cellular base station. The transmitter operates in a frequency band near that of the receiver and shares the same duplex filter.

The SNR in a communication channel characterizes the quality with which information is transferred through the channel. The SNR also sets a limit on the minimum required information carrying characteristics of the signal such that this information can be properly detected and recovered. By reducing the noise in a communication channel, the information can be communicated with higher fidelity; reducing the chances of being misunderstood or making errors. In digital communication, the performance is characterized as bit error rate (BER). The primary requirement of the communication system is to achieve and maintain a low BER by maintaining a very high SNR.

The communication channel considered in this thesis is the RF path present from the transmit power amplifier to the antenna of a wireless base station radio front-end. The power spectrum at the output of the transmit power amplifier contains the desired spectral components which fall within the transmit band of the communication channel used. Referring to **Figure 1.2.1**, it is shown that when an RF signal is amplified to form the transmit signal, a significant amount of emissions are generated outside the transmit band. The emissions are due to the non-linearity and noise generated inside the power amplifier, and may appear as a “skirt” or “shoulders” when observing the power spectrum at the output of power amplifier. In a dual band base station, the operating frequency bands of the receiver and transmitter are commonly separated by a fraction of the operating bandwidth. The closer the transmit frequency band is to the receive frequency band, the more emissions power is introduced in the receiver. The undesired spectral

components must be reduced to avoid the introduction of excessive noise in the receiver to increase the SNR. A duplex filter (duplexer) is used to provide sufficient isolation between the transmitter and the receiver.

Figure 1.2.1 – Power Amplifier Emissions Overlap with the Receive Band and Interfere with the Receive Signal at the Antenna Port of a Wireless Base Station.

A base station duplex filter typically consists of a number of precision machined microwave resonant cavities. The number of cascaded resonant cavities required depends on the required isolation specification of the transmit path filter and the band separation between the transmit and receive frequency bands. Increasing the isolation specification while reducing the transmit and receive band separation requires a very sharp transition band in the filter characteristic. Transition band sharpness can be achieved by cascading a number of resonant cavities to obtain the necessary frequency roll-off. The size of the duplex filter is hence proportional to the number of filter poles. Other size factors include power handling capabilities and materials used. Precise tuning of the resonant cavities is required before deployment. Reducing the size of the duplex filter and introducing a degree of frequency agility are currently some of the main challenges to overcome.

The function of the duplex filter is to provide isolation between the transmit signals and receive signals thus facilitating the use of one antenna per radio front-end module. The transmit channel bandpass filter provides attenuation in the order of 90 dB in the receive band to minimize the interfering noise emissions that degrades the performance of the receiver. The receive channel bandpass filter provides high attenuation in the transmit band to protect the sensitive receiver circuitry from high power signals that desensitize the receiver and could potentially saturate or damage the LNA. The duplex filter also assures that the radiated emissions are within specifications and reduces out-of-band noise in the receive path reducing the noise bandwidth. Reducing the noise bandwidth increases the receiver SNR.

The duplex filter specifications are derived from the emissions specifications established by the frequency spectrum regulating bodies, standard performance specifications established by the communication standards associations as well as some technological issues to be considered such as minimum level in power amplifier IMDs, PA noise floor and receiver sensitivity.

The filter market is currently dominated by personal communication systems (PCS) operating in the 800 MHz to 2 GHz frequency range. The duplex filter is the single largest component of the RF front-end module in the base station. The duplex filter occupies a significant portion of the space available in a rack-mount case in which the complete base station systems are installed; roughly 15 % of the volume of a typical base station. A typical duplex filter has dimensions of 14x8x3 inches and weights 20 lb.

Considering the case of a typical 800 MHz CDMA base station radio front-end, a full-duplex dual-band system, the 25 MHz receive band ranges from 824 MHz to 849 MHz and the 25 MHz transmit band ranges from 869 MHz to 894 MHz. The unused band of frequencies between the receive and transmit bands, ranging from 849 MHz to 869 MHz is the 20 MHz separation band. The bandwidth of the receive and transmit band occupy roughly 3% of the center frequency with a 2.3% bandwidth separation.

Figure 1.2.2 (a) and **(b)** shows typical receive channel (RX) and transmit channel (TX) passband specifications (spectral mask) for a cellular base station duplex filter operating in a system compliant to an 800 MHz CDMA communication standard [1]. In **Figure 1.2.2 (a)**, the S21 characteristics shows a minimum transmit band insertion loss of 85 dB. High power signals present at the PA output are reduced by 85 dB as seen by the receiver. The lower power IMDs at the output of the power amplifier are reduced in the separation band. In **Figure 1.2.2 (b)**, the general inverse S21 characteristic desired for the transmit path filter is shown. The 85 dB of receive band insertion loss reduces the out-of-band noise emissions from the PA that overlap in the receive band, effectively lowering the noise floor in the receive band to achieve performance specifications. The maximum transmit passband insertion loss in a typical duplex filter would be 0.65 dB.

Figure 1.2.2 – CDMA 800 MHz Duplex filter Passband Specifications

Assuming a typical scenario where the base station is operated at a PA average output power level of +35 dBm. Current linearized multi-carrier power amplifiers have typical IMD power levels approaching -15 dBm (-50dBc) in out-of-band emissions. The noise power level that consists mainly of noise emissions from the PA over the receive band can be assumed in the proximity of -26 dBm. The transmit path filter following the PA (transmit channel of the duplex filter) drops the receive band noise level by 85 dB to

achieve a receive band noise floor near -111 dBm. The power level in the transmit band as seen by the receiver is at a power level near -50 dBm where it does not significantly affect the dynamic range of the receiver. The noise in the receive band is sufficiently low to assure acceptable receiver SNR and avoid masking of the receive signal.

In theory, it is possible to use a feedforward architecture [2] or interference cancellation system to sample the undesired spectral components that interfere with the receive band. The sampled noise can then be used to cancel by vector addition, the unwanted noise generated by the PA before the interference enters the receive channel or degrades surrounding collocated receivers operating in the same frequency band.

1.3 – Review of Existing Active Noise Reduction and Interference Cancellation Technology Options

There are many ways to filter undesired spectral components. The choice of technology may depend on cost, manufacturability, power requirements and frequency band of operation [3]. Most commercial market manufacturers use coaxial combline filter technology to design duplexers operating in the GSM, CDMA and UMTS bands. The duplexers used are passive filters with Q-factors upwards of 10,000, power handling capabilities in the range of 100 W to 500 W and passband insertion losses below 1 dB. However, these filters require precise manual tuning to achieve specification and vary in frequency-selectivity over varying environmental conditions. Reducing the cost of mass fabrication of duplexers is a significant challenge.

Since the main function of the duplexer is to provide signal isolation from the transmit path to the receive path, the focus is on relaxing the specifications of the duplexer while maintaining sufficient isolation to assure no receiver sensitivity degradation. Relaxing the specification requirements of the duplexer would ultimately reduce its cost.

In developing a complementary noise reduction system for the duplexer unit, a few architectural and technical aspects must be kept in mind. The first aspect is simplicity of implementation and integration with current technology without affecting radio front-end components other than the duplexer. Hence, the signal paths to be considered should be restricted to the transmit path after the PA leading to the TX/RX antenna and the receive path before the receiver's LNA.

The next aspect to be realized is that the insertion loss in the RF path from the PA to the antenna must be kept at a minimum, below a maximum of 0.65 dB is typical in current 800 MHz CDMA base station duplexers. Insertion loss in the transmit path reduces the output power of the PA which is already working at its operating limit. Insertion loss introduced in the receive path directly impacts the noise figure. In a typical 800 MHz CDMA duplexer, the noise figure in the receive path must be lower than 2.2 dB after the LNA. Having any device in the path to the antenna would introduce insertion loss in both transmit and receive paths and should therefore be avoided. A non-intrusive method of ridding the noise in the transmit path would be best. Since insertion loss is unavoidable, it is best to concentrate on a solution that would only affect one path, minimally.

Active noise reduction techniques were investigated as a solution to complement the duplexer. Ivanov et al. in [2] present a review of feedback and feedforward noise reduction techniques. After some in-depth review of the problem, feedback was disregarded as an option due to its stability problems over broader bandwidths rendering feedback an inherently narrowband technique. A technique that seemed an interesting candidate solution to pursue this research was feedforward noise reduction, a result of feedforward cancellation. The technique also falls under the designation of interferometric carrier suppression, as a result of microwave interferometry. The technique works on the principle of vector addition of two microwave signals with the same amplitude but opposite phase. A microwave interferometric network is formed when a signal path is split in two at a junction and travels into two separate branches and

then is recombined at another junction. Designating the signal in one branch where a given process occurs (delay, amplification or other) as a reference, the signal traveling in the other branch, effectively by-passing the process, can be said to be “fed-forward” from the split point to the point where the signals are recombined.

Figure 1.3.1 – Constructive and Destructive Interference Between Two Signals

The technique based on the principle of superposition, makes it theoretically possible to selectively remove any spectral component of a signal from its RF spectrum. The cancellation technique, as applied to RF signals, will be referred to as feedforward cancellation for the remainder of this thesis. The principle is illustrated in **Figure 1.3.1** where a main signal and a cancellation signal arrive at a summing junction. If the two signals are in perfect anti-phase and have the same amplitude, as shown in **Figure 1.3.1 (b)**, no output results. Otherwise they add-up, as shown in **Figure 1.3.1 (a)**. This principle is well known as destructive and constructive interference of two waves.

Feedforward is becoming a very popular method for base station amplifier linearization [4]. Feedforward can theoretically provide perfect cancellation for all distortion signals, is modulation independent, inherently stable, capable of wideband operation and is well suited for dynamic channel allocation such as in cellular systems. The term “microwave interferometry” is used in the literature for near real time noise measurement system

(NMS) with an uncertainty limited primarily by the thermal noise in the lossy components of the microwave interferometer [5, 6].

Technological aspect of modern wireless standards can be harnessed to allow some degrees of freedom to approach the duplexer problem at the radio front-end level. The receive band is divided in many individual communication channels. The individual communication channels may overlap with each other. The base station receiver only listens to a portion of the receive band at any one time, and may be tuned from channel to channel as required. It is typical for wireless network operators to obtain license for a few channels and deploy only a few adjacent frequency bands in a service area. For example, a network operator may only have a license for a 5 MHz band in the allocated 25 MHz band. Having a filter that can be tuned actively to follow the receiver's listening channel would increase the signal to noise ratio in just the necessary channel being received. The concept is analogous to a frequency agile active duplexer. Feedforward is a good candidate for dynamic channel allocation since it can provide a very deep notch under narrow band conditions and can be electronically tuned to adapt to new conditions.

The next section gives some background information on the feedforward cancellation concept as well as some adaptive control schemes that can be used to implement optimization and frequency agility features.

1.3.1 – Feedforward Cancellation Technology

The principle of feedforward cancellation touches a few different fields of research and is the basis behind a few technologies. Feedforward cancellation has been widely employed recently as a linearization technique for power amplifiers [7, 8, 9, 10], low noise amplifiers [11] and other applications where ultra-linear amplifiers are required. Some of the most sensitive noise measurement systems [5, 6] and frequency discriminators use this technique. Feedforward cancellation has also been used to reduce flicker noise [12, 13] and design ultra-low noise in microwave oscillators that have

combined excellent phase and amplitude noise performance [14]. Frequency discriminators have been implemented using the technology [15, 16]. At much lower frequencies, active noise cancellation is well covered in the literature for applications such as active acoustic noise suppression in ventilation ducts and audio systems [17].

Feedforward cancellation concepts are also applied to interference cancellation systems to solve the problem of feed through that limits the dynamic range in radar systems [18]. Interference cancellation is not a new technology, it has been applied and driven by military applications in the past. R. N. Ghose and W. A. Sauter obtained a patent in 1972 for an adaptive interference cancellation system that can be used to ease collocation of high power transmitters and receivers [19, 20]. The same type of system has been suggested for use on electromagnetically dense platforms such as ships and aircrafts that contain many collocated transmitters and receivers [21]. In these systems, the interfering signal is sampled using an auxiliary antenna oriented in the direction of arrival of the interference signal or directly from the source such as the output of the power transmitter. The sampled signal is then used to subtract the interference present in the main antenna path of the affected receiver(s) [22].

A few companies specialize in interference cancellation systems, namely EDO Corporation/American Nucleonics (EDO ANC) [23], Adaptive Concepts, Inc. [24] and Cyber Dynamics Corporation [25], just to name a few. Although they have systems with outstanding performance, their production is not oriented towards high volumes of compact and cost effective solutions competitive in commercial telecommunication markets. As far as the author is aware at the time of this writing, this technique has never been applied to modern base station radio front-ends operating in the 800MHz CDMA and 2 GHz UMTS personal communication cellular markets.

Feedforward techniques are inherently unconditionally stable, have a theoretically infinite cancellation performance and can operate over wide bandwidths. Feedforward is

effective for multi-carrier signals such as CDMA and UMTS that have varying envelopes. The cancellation performance is independent of modulation type.

The main challenge behind feedforward techniques is the precision control of phase error, amplitude balance and time delay match between the cancelation signal and the main signal, required to achieve a level of cancellation performance in excess of 40 dB over a bandwidth of 25 MHz. Analysis of cancellation performance with respect to phase error, amplitude imbalance and delay mismatch is well covered in the literature. [10, 26, 27, 28, 29]. For example, to achieve a cancellation level of 30 dB, a phase error below 1° and an amplitude mismatch below 0.1 dB must be maintained at a given instantaneous frequency. The cancellation bandwidth is improved by controlling the delay match between the two paths by minimizing phase and group delay difference between the two paths over a given bandwidth.

1.3.2 – Adaptive Control Systems

The balance between the two paths of a microwave interferometric network tuned to reject a given frequency band in a typical cancellation system is sensitive to disturbances such as environmental fluctuations (temperature, stress), aging, change in operating center frequency and input power level variations. The balance between the two branches of the microwave interferometric network may be maintained for short periods of time ranging from a few minutes to a few hours but is likely to vary over longer periods of time. Disturbances cause phase and amplitude imbalances, which degrade the cancellation performance.

The most basic type of feedforward architecture has a signal controller in one of the two branches forming the microwave interferometric network. The signal controller is the core of the noise cancellation system. The signal controller must be able to produce a counter-interference signal having a wide instantaneous bandwidth and wide dynamic range. The main component in the signal controller could be a vector modulator, a

cascade of an amplitude attenuator and phase shifter or other component capable of modifying the relative amplitude and phase of the signal passing through. An amplifier may also be contained in the signal controller in the case where there is too much loss in the feedforward path to match the amplitude of the interference signal present in the main path. In a basic feedforward cancellation system the delay is matched at the point of implementation by tuning a delay line manually and is not controlled after deployment since the system tolerance to disturbances is usually reasonable to maintain the required bandwidth performance. However, the signal controller could also contain some means of fine tuning the electrical delay of the RF signal [30].

To keep the system optimally balanced, some form of adaptive control system must be used. The adaptive control systems senses the current performance of the system and by extracting the information necessary to improve the performance, it adapts to new operating conditions and tracks the optimal cancellation level. Adaptive control systems can also be used to implement desirable features such as frequency agility.

Many patents have been issued relating to inventions that improve the cancellation performance of feedforward cancellation systems by addressing a particular problem or controlling specific variables. Many techniques have been used to adaptively modify the cancellation signal [31, 32, 8, 32, 35]. There are various approaches to the control problem of a feedforward cancellation loop. Most popular approaches generate an error signal from the residual cancellation vector present at the output of the summing junction (point of cancellation). The information extracted from the error signal depends on the choice of control algorithm. Some control algorithms are implemented in DSP and attempt to minimize the cancellation residual power level. Such DSP algorithms may only detect the average power or the signal envelope using simple diode detectors. On the other hand, a control scheme that makes use of a pilot tone injected in the system makes use of a synchronous receiver to reconstitute the error vector and make necessary adjustment to minimize the error vector by either controlling the phase and amplitude (polar loop techniques) or the I and Q components (Cartesian loop techniques). In the

Cartesian approach, the cancellation residual vector is synchronously correlated with a reference pilot tone signal and appropriate error signals are generated from the synchronous detector. Other methods can make use of the input signal information and a digital lookup table to adjust the cancellation vector [36].

The control challenge is to implement a control system that operates to minimize the cancellation residual such that the spectral mask requirements for the application are met. The control system must maintain the specified residual level in the event that the frequency at which the cancellation is applied, is changed.

1.4 – Noise Cancellation Technology Performance Comparison Matrix

Some results available in the literature that use feedforward cancellation methods were obtained and compiled in **Table 1.4.1**. The table contains the measured cancellation performance of different feedforward cancellation systems implemented in hardware. Most of these systems were applied to linearized amplifiers. It is expected that very high performance interference cancellation systems are still classified, proprietary or very expensive. The focus has been on implementations of low to moderate complexity that would prove suitable for commercial use in the telecommunication industry. Some companies claim that over 60 dB of cancellation is achievable over wide bandwidths in the frequency range of 2 MHz – 12 GHz, however technology specific details have not been found.

A few interesting points are worth mention. In [37], phase equalizers were used instead of pure delay lines to deliberately introduce deviation from linear phase (DLP) characteristics with frequency in the delay path to approximate those of an amplifier contained in the main path. This improved the cancellation performance over a wide band by reducing the frequency dependent phase shift occurring at different frequencies; essentially equalizing the group delay characteristic of the two paths.

In [38], a wide band design (25% center frequency) maintained a cancellation of 20 dB while a narrow band design (1% of center frequency) has achieved 55 dB of cancellation. The authors claim that better bandwidth and higher cancellation performance can be obtained by improving the return loss between the subcircuits.

Table 1.4.1 - Feedforward Cancellation Performance Comparison						
Cancellation (dB)	Bandwidth (MHz)	Center Frequency (MHz)	Application	Notes / Technology	Reference	Year
30	15	1,475	SAFF- PA Multi-Carrier Amp	Pilot Tone + Correlator, DSP Controller	[8]	1991
50	30	880	Broadband PA		[39]	1991
20	500	6,200	Broadband Feedforward Amplifier		[41]	1996
40 / 30	7 / 10	815	Feedforward Amplifier	DSP Adaptive Control	[17]	1996
35	30	1,855	Linear PA		[28]	1997
24 / 55	200 / 18	1,800	Feedforward Amplifier	Imped. & Isolat. mismatch Analysis	[38]	1998
25	80	1,800	Wideband Feedforward Amplifier	DSP Adaptive Control	[42]	1998
40 / 25	40 / 80	1,800	FFPA	2-Branch Compensator	[42]	1998
24	20	450	PA Linearization	Adaptive Control (Polar)	[9]	1999
30	195	2,107	Feedforward Amplifier	Used variable mechanical capacitors for phase tuning	[40]	1999
20	200	1,800	Wide-band Feedforward Amplifier	Use of Phase Equalizer	[37]	2001
20	25	836	Feedforward Low Noise Amplifier		[11]	2001
50 / 40 / 30	4.5 / 14.4 / 45	836.5	FFCS Feedforward Wideband (25 MHz) Notch Filter	Manual Tuning of I/Q vector Modulator	This Thesis	2003
25 / 30	4.5 / 14.4	836.5	FFCS Feedforward Wideband (25 MHz) Notch Filter, Frequency Agile	Analog Adaptive Control (I/Q VM and DM) using Pilot Tone control signal (** non-optimized control loop)	This Thesis	2003

In [39], a remarkable 50 dB of cancellation has been reported in a 3.5% bandwidth centered at 880 MHz. There was no mention on the type of control or tuning method to adjust the phase and amplitude of delay paths. In [40], 30 dB of cancellation is claimed over a 9.25% effective bandwidth.

From the results, it can be observed that 30-40 dB of cancellation is practically feasible over 3% bandwidths. Most of the hardware prototypes tested were designed for experimental purposes and were not fine tuned or designed in budgets allowed for commercial products and high volume manufacturability. It should also be noted that in most reported results, cancellation of intermodulation products of linearized amplifiers was reported. The intermodulation products at the output power amplifiers are higher near the carrier frequencies and decrease at frequencies further from the carriers or test tones. The cancellation characteristic “V”-shape may have been appropriate to drop all intermodulation products below the thermal noise floor of the measuring instruments at the settings used. Therefore, it is not clear whether all the reported cancellation performances are effective over the entire bandwidths of operation. Furthermore, cancellation performance of the carrier loop is usually reported better than the intermodulation cancellation loop since carrier cancellation is a narrow band process compared with the wider-band process of canceling the intermodulation products over the entire output spectrum of the amplifier.

1.5 – Target Performance Requirements of Proposed Noise Cancellation System

The objective of this research was to cancel the IMD power spectral components generated by the PA, that fall in the receive band, by a level of 40 dB. By reducing the noise level in the receive band before the duplexer, the specifications of the duplexer can ultimately be relaxed. The solution was to use feedforward noise cancellation to achieve the objective. A variation on prior-art adaptive interference cancellation system and feedforward noise reduction technique is proposed as a solution.

It was decided to implement the first prototype to operate in the 800MHz CDMA band. The choice of technology was based on ease of implementation, avoiding potential parasitic effects arising with technologies at shorter wavelength that would fall outside the scope of this research and wide selection of high-performance discrete surface mount components readily available on the market in the 800 MHz CDMA band.

The objectives set for the research and prototype implementation were:

- Develop a proof of concept for a novel adaptive control system that would allow the feedforward interference cancellation system (FFCS) to operate in a bandwidth of 25 MHz in the CDMA receive band (824 MHz – 849 MHz).
- The FFCS should introduce a minimum of 30 dB of rejection in an instantaneous 5MHz bandwidth and be tunable anywhere in the 25 MHz CDMA receive band.¹
- The FFCS should introduce a maximum pass band insertion loss of 0.65 dB in the transmit band (869 MHz – 894 MHz).
- The FFCS should be capable of handling +45 dBm signals in the transmit band.
- The FFCS should be easily integrable with the existing base station hardware.

In an actual product implementation, the FFCS would be able to achieve a 30 dB reduction in the level of out-of-band transmitted noise and IMD products originating from the PA. Without the FFCS, the noise generated by the PA would travel through the transmit-receive junction port to the antenna and appear in the receive band of the CDMA base station radio front-end receiver. Combined with a more compact, less expensive and reduced order duplexer, the (receive band + transmit noise) power level of -30 dBm would be reduced by 85 dB at the transmit-receive junction to the antenna port. To be useful for typical wireless network deployments, the system should be effective over an instantaneous bandwidth of 5 MHz located anywhere in the 25 MHz CDMA receive band which is centered at 836.5 MHz. A successful product using the FFCS technology would ultimately reduced the cost of current base station while maintaining or exceeding the existing duplex filter performance specifications.

¹ A similar approach to the control problem used in this thesis to allow frequency agility to a tunable resonator has emerged in the literature near the end of this research [43].

Specialized components with very tight tolerances or complex adaptive control circuit are required to satisfy the requirements to apply the concept to duplexer technology. It is essential to understand the feedforward cancellation method both at a theoretical and practical level to make effective use of this technology.

1.6 – Conclusion

From the literature review, it was concluded that a feedforward noise cancellation system (FFCS) could be developed to complement a duplex filter of reduced order, reduced size and reduced cost. The device, combined with the duplexer would fit the performance specifications for duplexer characteristics operating with modern and future cellular technologies.

The FFCS should be able to achieve a reduction of 30 dB in the -26 dBm noise level generated by the PA that fall within the receive band that spans a bandwidth of 25 MHz centered at 836.5 MHz. The duplexer transmit channel insertion loss specification could be relaxed from 85 dB to 55 dB.

The result from relaxing the specifications of the duplexer would reduce the number of poles in the transmit filter from 7 to 4. One pole corresponds to one resonant cavity in the filter. The 800 MHz CDMA duplexer actually has 2 receive filters with 7 poles each. Removing 3 poles from the transmit filter characteristic would reduce the volume by 15 %. The price of the duplexer is proportional to the size of the duplexer. The reduced size duplexer would benefit a 15 % cost reduction. The added cost of the FFCS module would have to be less than the cost reduction to achieve a significant benefit. As technology is miniaturized, significant cost reduction may be achieved using this technology.

A potential benefit of the FFCS would be to render the duplexer frequency agile. A frequency agile duplexer could have an impact on future generation communication

standards by adding some flexibility in the spectral mask requirements. The spectral mask requirements for a base station may not be achieved over the entire receive band, however, the SNR requirements to meet the performance specification over the band of operation may be sufficient and dynamically tunable. Frequency agility could have great potential, allowing wireless networks to be reconfigured to operate at a different frequency band without modifying the actual hardware in the field.

Chapter 2 – Theoretical Analysis of Fundamental Noise Cancellation Level

Determining Factors and Simulation Results of the FFCS Architecture

2.0 - Introduction

It is essential for effective design to understand the key mechanisms that determine the cancellation level of the FFCS. The basic feedforward cancellation topology depicted in **Figure 2.0.1** is used to derive the theory of operation of feedforward cancellation and serve as a system model for preliminary system level simulations.

Figure 2.0.1 – Basic Feedforward Cancellation Topology

In the context of **Figure 2.0.1**, the basic feedforward cancellation process consists of adding the anti-phase cancellation signal S_{can} to the main signal S_{main} . The signals S_{main} and S_{can} are input to the cancellation coupler in which is located the point of cancellation. In the simplest case where the signal S_{can} is equal to the signal S_{main} , the signal in the noise band at the output of the cancellation coupler has zero RMS power. The signal components of S_{main} have been perfectly cancelled reducing the noise floor of the main path to that of the cancellation path over the cancellation bandwidth.

In a practical application of feedforward cancellation technology, an undesired spectral components S_n is present in the power spectrum containing a desired RF signal S_i . S_n commonly consist of intermodulation distortion components (IMD) introduced by an RF amplifier or artificial noise source that is picked up or induced in the signal path. In a typical communication system, S_n reduces the signal to noise ratio (SNR) at the receiver input of the system and ultimately reduces the sensitivity of system's RF receiver. S_n must be removed to improve the SNR of the received signal.

As opposed to an amplifier that is linearized using feedforward technology, the application for the feedforward noise cancellation system developed is to perform as a band-stop filter with an optimized rejection notch. The difference between the two applications is that in a feedforward linearized amplifier, the cancellation bandwidth is theoretically infinite, since no frequency selectivity in the cancellation process is required. On the other hand, the FFCS targets the cancellation of specific frequency components already introduced in a signal path propagating along with a desired signal. The desired signal must not be disturbed by the cancellation process.

In a feedforward cancellation system, the noise signal S_n is sampled. In practice, the process of sampling the noise signal and injecting the cancellation signal in the main path introduces loss, phase shift and delay of the cancellation signal S_{can} relative to the noise signal portion S_n present in the main path (see **Section 2.2.1**). The feedforward cancellation system must modify the sampled noise signal S_{noise} to reproduce an anti-phase replica of S_n at the point of cancellation. The signal processor adjusts S_{noise} to produce S_{can} . Coupling loss and phase shift incurred by S_{noise} is introduced by the signal sampling and signal injection directional couplers. The coupling losses and other losses in the cancellation path are overcome by amplifying the sampled noise signal by a gain that will assure amplitude balance between the S_n and S_{can} at the point of cancellation. The amplifier introduces further phase shift, and the gain cannot necessarily be balanced exactly as required in practice. The amplified S_{noise} is then fine-tuned by a vector

modulator such that its phase and amplitude are ideally the perfect inverse of S_n at the point of cancellation.

The cancellation signal is ultimately delayed by propagating through the different stages of the signal processor. The signal processing delay can be expressed as the electrical time delay through the cancellation path from the point where the signal is sampled to the point of cancellation. The delay must be compensated in the main path with a delay line τ_{main} introduced between the point where S_n is sampled and the point of cancellation. The cancellation signal S_{can} , which is the inverse of S_n , is then re-injected in the main path to cancel the undesired spectral components present in the composite main path signal. A successful cancellation process reduces the spectral components in S_n to zero while maintaining the integrity of the desired signal S_i .

The integrity of the main path signal is preserved by introducing minimal insertion loss and minimal distortion by sampling only the noise signal. Sampling only the noise signal S_n assures that the injected cancellation signal is only correlated with the noise signal in the main path. Minimal insertion loss in the main path is achieved by sampling S_n and injecting S_{can} in the main path with directional couplers. In an application where the frequency bands of S_i and S_n are separated, S_n may be sampled selectively by introducing a bandpass filter with high rejection over the band of S_i in the cancellation path, just after the noise sampling coupler. Furthermore, narrowband directional couplers may only couple the power over the designed frequency range of operation, essentially minimizing the introduction of S_i components in the cancellation path. The SNR out of the FFCS in the transmit band should be not be degraded when compared to the SNR into the FFCS in the transmit band (negligible noise figure). On the other hand, the SNR into the FFCS over the receive band should be significantly higher than the SNR out of the FFCS over the receive band. The noise figure in the transmit band should be equal to the insertion loss of the main path while the noise figure in the receive band should be negative.

To simplify analysis, the main signal S_{main} considered in this chapter is assumed to be composed entirely of noise components S_n to be removed. There are no desired signal components S_i in S_{main} that need to be preserved.

Phase error, amplitude imbalance and delay mismatch between the main signal and cancellation signal, are the fundamental parameters that determine to a first order the level of cancellation that can be achieved at the point of cancellation. The level of cancellation achieved is a measure of the performance of the feedforward cancellation system.

The phase error is the difference in phase between S_{main} and S_{can} . The amplitude imbalance is the difference in amplitude between S_{main} and S_{can} . The delay mismatch is the difference in the time of arrival of S_{main} and S_{can} at the point of cancellation where actual signal interference takes place. With proper signal processing, the phase error, amplitude imbalance and delay mismatch can all be minimized, resulting in maximum cancellation level.

Theoretical analysis of the impact of phase error, amplitude balance and delay error on the achievable cancellation level is investigated in the following sections. The theoretical calculations to generate the plots and analyze the cancellation level are performed in Mathcad. Mathsoft's Mathcad Professional is a simple and powerful computer software mathematical analysis tool ideal for performing engineering calculations.

The derived theoretical analysis and system simulation results obtained in this chapter serve as a benchmark to assess the hardware performance of the feedforward cancellation prototype and provide an initial system level proof of concept for the prototype to be implemented in hardware.

2.1 –Analysis of the fundamental cancellation level determining factors

For simplicity of analysis, the main signal and cancellation signal are assumed to be two single tone sinusoidal signals. Equations (2.1.1) and (2.1.2) characterize the main and cancellation path instantaneous voltage vectors at the point of cancellation. Although this simplification is specific to the case of two sinusoidal signals, it will be directly applicable to initial characterization of the hardware implementation developed since the system makes use of a pilot tone for centering the cancellation at a specific frequency.

$$v_{\text{main}}(t) := V_{\text{main}} \cdot \cos(2 \cdot \pi \cdot f_c \cdot t - \beta_c \cdot l_{\text{main}} + \phi_{\text{main}}) \quad (2.1.1)$$

$$v_{\text{can}}(t) := V_{\text{can}} \cdot \cos(2 \cdot \pi \cdot f_c \cdot t - \beta_c \cdot l_{\text{can}} + \phi_{\text{can}}) \quad (2.1.2)$$

V_{main} is the amplitude of the main signal. V_{can} is the amplitude of the cancellation signal. f_c is the instantaneous frequency (considered constant) of both signals. β_c is the phase shift constant at f_c for a particular transmission line. ϕ_{main} and ϕ_{can} are the initial phase of the main signal and cancellation signal respectively. l_{main} and l_{can} are the length of transmission line that each signal passes through before being input to the cancellation coupler.

The vector addition of two signals in equations (2.1.1) and (2.1.2) results in an error component characteristic of imperfect cancellation. If the cancellation system is perfect, the error signal is ideally reduced to zero. The magnitude and phase characteristics of the error signal depend on the phase error, amplitude imbalance and delay mismatch between the main signal and cancellation signal at the point where the two signal interfere. Equation (2.1.3) redefines the cancellation signal of equation (2.1.2) as a function of the three fundamental cancellation level determining factors taking the main signal as a reference.

$$v_{\text{can}}(t) := (V_{\text{main}} + V_{\text{diff}}) \cdot \cos[2 \cdot \pi \cdot f_c \cdot t - \beta_c \cdot (l_{\text{main}} + \Delta l) + 180^\circ + (\phi_{\text{main}} + \theta_{\text{err}})] \quad (2.1.3)$$

$$\theta_{\text{err}} := (\phi_{\text{can}} - \phi_{\text{main}}) - 180^\circ \quad (2.1.4)$$

$$V_{\text{diff}} := V_{\text{can}} - V_{\text{main}} \quad (2.1.5)$$

$$\Delta l := l_{\text{can}} - l_{\text{main}} \quad (2.1.6)$$

In equation (2.1.3), θ_{err} is the phase error between the main signal and the cancellation signal, defined by equation (2.1.4). V_{diff} is the voltage amplitude difference between the main signal and the cancellation signal, defined by equation (2.1.5). Δl is the difference in transmission line length resulting in an effective delay mismatch between the main signal and cancellation signal, defined by equation (2.1.6).

The error signal defined by equation (2.1.7) is the result of the vector addition of the main signal and the cancellation signal.

$$v_{\text{err}}(t) := v_{\text{main}}(t) + v_{\text{can}}(t) \quad (2.1.7)$$

In RF communications systems, it is common practice to work with power levels to characterize signals and compare signals in terms of power ratios instead of working with the signal voltage levels. To introduce the concept of cancellation performance, the main signal and error signal are expressed in terms of their average power levels at the point of cancellation in equations (2.1.8) and (2.1.9). Z_0 is the characteristic impedance of the system where signals are present (usually 50 Ω) and T_c is the time period of the sinusoids.

$$P_{\text{main}} := \frac{1}{T_c} \int_0^{T_c} \frac{v_{\text{main}}(t)^2}{Z_0} dt \quad (2.1.8)$$

$$P_{\text{err}} := \frac{1}{T_c} \cdot \int_0^{T_c} \frac{v_{\text{err}}(t)^2}{Z_o} dt \quad (2.1.9)$$

The cancellation performance of a cancellation system is a measure of the cancellation level by which the main signal power level in a certain bandwidth is reduced. The cancellation level **CP** achieved is defined as the power ratio of the residual error signal to the main signal at the point of cancellation. The cancellation level **CP** is expressed in units of decibels (dB) in **(2.1.10)**.

$$CP := 10 \cdot \log \left(\frac{P_{\text{err}}}{P_{\text{main}}} \right) \quad (2.1.10)$$

The fundamental cancellation level determining factors and their contribution to cancellation performance are studied further in the following sections. The determining factors are first considered individually and then collectively to indicate the relative sensitivity of the cancellation level to each factor.

2.1.1 –Cancellation Performance as a Function of Amplitude Imbalance Only

The effect of amplitude imbalance between the main signal and the cancellation signal on cancellation performance is considered assuming there is no phase error and no delay mismatch between the two signals. The amplitude imbalance is defined in equation **(2.1.5)** as the difference in voltage level between the main signal and the cancellation signal. Amplitude imbalance can also be expressed as the voltage amplitude difference between the cancellation signal and the main signal normalized to the main signal voltage. The normalized amplitude imbalance is defined in equation **(2.1.1.1)**. It is convenient to express the normalized amplitude imbalance in units of decibels using equation **(2.1.1.2)**.

$$\alpha := \frac{V_{\text{main}} + V_{\text{diff}}}{V_{\text{main}}} \quad (2.1.1.1)$$

$$\alpha_{\text{dB}} := 20 \cdot \log(\alpha) \quad (2.1.1.2)$$

Following the theory from **Section 2.1**, the closed form expression for cancellation performance as a function of amplitude imbalance in units of (dB) becomes equation **(2.1.1.3)** (see **Appendix A1** for derivation).

$$\text{CP}(\alpha_{\text{dB}}) := 10 \log \left(1 + 10^{\frac{2 \alpha_{\text{dB}}}{20}} - 2 \cdot 10^{\frac{\alpha_{\text{dB}}}{20}} \right) \quad (2.1.1.3)$$

Figure 2.1.1.1 – Cancellation Performance as a Function of Amplitude Imbalance

The cancellation performance is plotted as a function of amplitude imbalance in **Figure 2.1.1.1**. The vertical scale in **Figure 2.1.1.1** shows the level by which the main signal would be attenuated coming out of the point of cancellation for different levels of amplitude imbalance between the main signal and the cancellation signal. When the amplitude of the cancellation signal exceeds that of the main signal, V_{diff} is positive and the amplitude imbalance is positive. The solid trace represents the cancellation performance as a function of positive amplitude imbalance. When the amplitude of the main signal is larger than that of the cancellation signal, V_{diff} is negative, hence the amplitude imbalance is negative. At one extreme, the cancellation signal can be zero, causing the cancellation performance to be zero (i.e. S_{main} is unchanged). At the other extreme, S_{can} can be amplified to point where it is much larger than S_{main} causing the cancellation performance to grow in the positive direction (i.e. the output of the cancellation process has more power than S_{main}). The circle marker on the plot shows that for a cancellation level better than 30dB, the amplitude imbalance must remain below $\alpha_{\text{dB}} = 0.270$ dB (or equivalently above $\alpha_{\text{dB}} = -0.279$ dB).

2.1.2 –Cancellation Performance as a Function of Phase Error Only

The effect of phase error between the main signal and the cancellation signal on cancellation performance is considered assuming no amplitude imbalance and no delay mismatch. Phase error is expressed as the phase deviation from perfect antiphase (hence the -180° term) between the cancellation signal and the main signal as defined in equation (2.1.2.1). Phase error is expressed in units of degrees ($^\circ$).

Following the theory from **Section 2.1**, the closed form expression for cancellation performance as a function of phase error in units of (dB) becomes equation (2.1.2.1) (see **Appendix A2** for derivation).

$$\text{CP}(\theta_{\text{err}}) := 10 \cdot \log \left[2 \left(1 - \cos(\theta_{\text{err}}) \right) \right] \quad (2.1.2.1)$$

The cancellation performance as a function of phase error is plotted in **Figure 2.1.2.1**.

The figure demonstrates the cancellation performance for different degrees of phase error between the cancellation signal and the main signal. When the phase error is between 90° and 270° , the two signals interfere constructively and there is no cancellation. Constructive interference results in amplification of the main signal. Destructive interference only occurs at phase errors less than 90° and greater than 270° . At a cancellation performance of 0 dB, the main signals and cancellation signal are in quadrature. The characteristic indicates that there are two solutions for a given level of cancellation performance. The two solutions correspond to a positive or negative phase error value. The circle marker and square marker indicate that to achieve a cancellation performance better than -30 dB, the absolute phase error must be less than 1.8° and -1.8° respectively.

Figure 2.1.2.1 – Cancellation Performance as a Function of Phase Error

2.1.3 –Cancellation Performance as a Function of Amplitude Imbalance and Phase Error

The effect of both amplitude imbalance and phase error between the main signal and the cancellation signal on cancellation performance is considered assuming there is no delay mismatch. The closed form expression for cancellation performance as a function of both amplitude imbalance (in dB) and phase error becomes equation (2.1.3.1) and has units of (dB) (see **Appendix A3** for derivation).

$$CP(\alpha_{dB}, \theta_{err}) := 10 \log \left[1 + \left(10^{\frac{2\alpha_{dB}}{20}} \right) - 2 \cdot \left(10^{\frac{\alpha_{dB}}{20}} \right) \cdot \cos(\theta_{err}) \right] \quad (2.1.3.1)$$

Figure 2.1.3.1 – Cancellation Performance as a Function of both Amplitude Imbalance and Phase Imbalance

Figure 2.1.3.1 graphs the cancellation performance in presence of different levels of amplitude imbalance and varying degrees of phase error at the point of cancellation between the main signal and the cancellation signal. The phase error is plotted on the horizontal axis. Each curve corresponds to a different level of amplitude imbalance. The curves are in order from top to bottom and correspond one-to-one with the y-axis functions. The top curve starting at -12 dB of cancellation performance corresponds to an amplitude imbalance of 2 dB. The cancellation performance is limited by the combination of both factors. To achieve a cancellation performance better than -30 dB, the amplitude imbalance and phase error must simultaneously be less than 0.2 dB and 1.8° respectively, while they individually require 0.3 dB and 1.8° .

Figure 2.1.3.2 – Surface Plot of the Cancellation Performance as a Function of Amplitude Imbalance and Phase Error

The coupling between the amplitude imbalance and phase error is better represented by the cancellation performance surface. **Figure 2.1.3.2** illustrates a surface plot of the cancellation performance as a function of both amplitude and phase error. The outline of an XY-circle resulting from taking a horizontal slice of the vortex-shaped surface for a given level of cancellation performance yields all possible amplitude imbalance and phase error combinations that result in that level of cancellation. Any set of amplitude imbalance and phase error coordinate value inside the boundary set by the circle yields a better cancellation performance.

2.1.4 –Cancellation Performance as a Function of Delay Mismatch

The effect of delay mismatch between the main signal and the cancellation signal on cancellation performance is considered assuming perfect amplitude balance and no phase error. The cancellation path and the main path have different delay characteristics. The main reason why the two paths have different delay characteristics is because each path uses different components. While the only passive component in the main path is a delay line, the cancellation path contains both passive and active RF components, namely couplers, amplifiers and transmission lines. Each RF component has a particular phase delay characteristic and group delay characteristics. The delay characteristic from one component to another may be different. The mismatch between the delay characteristics of the main path and cancellation path create an important limitation on the cancellation performance.

The delay mismatch between the main path and the cancellation path is defined as the difference in arrival time between the two signals at the point of cancellation. The delay mismatch may also be interpreted as a phase difference or the difference in propagation wave length at a given frequency. The latter is chosen for the context of this thesis. The delay characteristic of each path is a function of both phase delay and group delay of the different components encountered by a signal in each path. Only phase delay characteristics are considered for the following analysis and zero group delay is assumed.

The group delay is the level of phase dispersion (deviation for linear phase) or non-linear phase shift characteristic as a function of frequency. A transmission line with group delay causes frequency components at different frequency (or wavelength) to arrive at different times at a given point along the transmission line, effectively distorting the wave-front arriving at that given point. In the context of a communication system, the effect of group delay blurs the information by reducing the correlation between the sent signal and the signal intended to be received through the propagation medium. The effect of group delay is mainly observed when considering dispersive channels with limited coherence bandwidths [44]. Considering the well controlled propagation mediums used over the bandwidth of operation of the FFCS, the effect of group delay can safely be assumed negligible for the first order analyses that follow. The delay characteristics are introduced by first defining the effective dielectric constant as a function of frequency of the transmission medium [45 p. 145-153].

The wavelength of a signal at a given frequency component f , propagating in a transmission medium characterized by a dielectric constant ϵ_r (which for microstrip would be ϵ_{eff}) is defined by equation (2.1.4.1) where c is the speed of light in free space. The wavelength has units of meters (m) and is inversely proportional to the frequency in a non-dispersive transmission medium. The phase constant β of the propagation medium is defined in equation (2.1.4.2). The phase constant is function of wavelength (frequency) and has units of radians per meter (rad/m).

$$\lambda(f) = \frac{c}{f \cdot \sqrt{\epsilon_r}} \quad (2.1.4.1)$$

$$\beta(f) = \frac{2 \cdot \pi}{\lambda(f)} \quad (2.1.4.2)$$

The cumulative phase shift (“unwrapped” phase) can be expressed as a function of the distance traveled along the transmission line. Assuming two paths have the same phase

constant (i.e. same phase velocity), the length difference Δl between the two paths can be expressed as the wavelength difference at the center frequency where the delay match is intended to occur. Equation (2.1.4.3) expresses the phase shift as a function of the frequency and the delay mismatch between the two paths. A signal component at a given frequency will experience a total phase shift of $\Delta\phi$ after traveling a distance Δl . The negative sign in the equation is used to indicate that the phase shift is a retardation phase, since it occurs at a later time relative to the reference phase at time zero, from the point where it departed.

$$\Delta\phi(f, \Delta l) = -\beta(f) \cdot \Delta l \quad (2.1.4.3)$$

The time delay is expressed in units of seconds (s) by dividing (2.1.4.3) by the radial frequency ω which is in units of (rad/s). Time delay is expressed in equation (2.1.4.4). The time delay is independent of frequency and is a function of the traveling distance in a given propagating medium.

$$\tau(\Delta l, f) = \frac{\beta(f) \cdot \Delta l}{\omega} = \frac{\beta(f) \cdot \Delta l}{2 \cdot \pi \cdot f} = \frac{\sqrt{\epsilon_r}}{c} \cdot \Delta l \quad (2.1.4.4)$$

Figure 2.1.4.1 illustrates the relative phase shift undergone by the main signal and cancellation signal at the point of cancellation as a function of normalized frequency. A fixed path length difference of Δl as defined in equation (2.1.6) is assumed between the two paths. The relative phase shift θ_{main} at the point of cancellation is shown as a solid line while the θ_{can} is shown as a dash line. In the case of **Figure 2.1.4.1**, zero phase error is assumed between the two signals at frequency component f_c . In presence of the phase mismatch, the slope of the two phase curves is different. Perfect phase match can only occur at f_c . Therefore, an effective phase error results from the delay mismatch at every other frequency than the one for which the two paths are matched for zero phase difference.

Figure 2.1.4.1 – Relative Phase Shift of Two Delay Mismatched RF Paths

The relative phase shift difference $\Delta\theta$ between the two paths may be expressed as a function of frequency. The dielectric constant of the two paths may be different, as commonly experienced when using a coax delay line to compensate for the delay through a microstrip RF path. A more intuitive and general way to qualify the delay mismatch between two non-dispersive paths is to express the delay mismatch as the difference in the number of wavelengths λ_{err} between the two paths at the design center frequency. For two paths with the same dielectric constant, the resulting number of wavelength difference would be $\lambda_{\text{err}} = \Delta l / \lambda_c$, where λ_c is the wavelength at the center frequency in the propagating medium. Subtracting θ_{main} from θ_{can} and expressing the result as a function of λ_{err} represents the relative phase shift difference as a function of frequency f with respect to the center frequency f_c , and is expressed in equation (2.1.4.5) (see **Appendix A4 a)** for derivation). At the center frequency, there is no relative phase shift error caused by the delay mismatch. Delay mismatch only introduces relative phase error at frequency components other than the center frequency.

$$\Delta\theta(f, \lambda_{\text{err}}) := 2\pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c}\right) \quad (2.1.4.5)$$

Figure 2.1.4.2 illustrates the relative phase shift difference between the two paths as a function of normalized frequency for two values of wavelength difference at the design

center frequency. For example, in presence of a 10 % wavelength difference ($\lambda_{\text{err}} = 0.1$), the phase error of 3.6° results between the two signals at 10 % of the center frequency.

Figure 2.1.4.2 – Relative Phase Shift Difference as a Function of Wavelength Difference

The cancellation performance can be expressed as a function of frequency f and delay mismatch λ_{err} by substituting $\Delta\theta(f, \lambda_{\text{err}})$ for θ_{err} in equation (2.1.2.1) to obtain the cancellation performance expressed in equation (2.1.4.6). The plot of the cancellation performance as a function of frequency and difference in the number of wavelengths with respect to the center frequency is shown in **Figure 2.1.4.3**.

$$CP(f, \lambda_{\text{err}}) := 10 \cdot \log \left[2 \cdot \left[1 - \cos \left[2 \pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c} \right) \right] \right] \right] \quad (2.1.4.6)$$

Figure 2.1.4.3 demonstrates the cancellation performance between the main signal and cancellation signal in presence of delay mismatch. To achieve a cancellation performance better than -30 dB in a 25 MHz bandwidth centered at $f_c = 836.5$ MHz requires the delay mismatch to be within $(1/3)$ of the center frequency wavelength. This is equivalent to a relative path phase difference of 120° or an electrical time delay of $\lambda_{\text{err}} / f_c = 398$ ps. The cancellation bandwidth at a given level of cancellation performance is therefore limited by delay mismatch.

Figure 2.1.4.3 – Cancellation Performance as a Function Delay Mismatch and Normalized Frequency

The case where multiple wavelengths difference is present between the two paths is also shown in **Figure 2.1.4.3**. Multiple cancellation nulls occur at frequency components where the relative phase shift difference between the two paths is a multiple of 2π , where the phase error becomes zero. The two signals add constructively at frequency components where the two signals experience a 180° relative phase difference. The reader should take note that in practice, the cancellation nulls at the offset frequency components where cancellation re-occurs, would be observed to be less pronounced as the frequency deviation from the center frequency gets larger. Such excessive difference in path length is prone to reduce the coherence bandwidth between the two signals. However, the path delays of a practical design can be matched to within fractions of wavelength at the center frequency. Also, a designer could purposely introduce a delay

mismatch, for example, to achieve an insertion loss goal by shortening the main path delay line at the expense of reduce cancellation performance.

Delay mismatch between the main signal and cancellation signal at the point of cancellation becomes an important limiting factor in the cancellation performance when a wideband signal is to be suppressed. Other practical issues such as temperature dependent characteristics of hardware components used render wideband delay match harder to achieve in practice due to amplitude ripple and phase error requirements.

2.1.5 – Cancellation Performance as a Function of all Fundamental Cancellation Determining Factors.

The effect of amplitude imbalance, phase error and delay mismatch between the main signal and the cancellation signal on cancellation performance are considered simultaneously. (2.1.4.5) is summed with θ_{err} in (2.1.3.1) to obtain the closed form expression for cancellation performance as a function of all fundamental cancellation performance determining factors. The cancellation performance function in units of (dB) is expressed in equation (2.1.5.1).

$$\text{CP} \left(\alpha_{\text{dB}}, \theta_{\text{err}}, \lambda_{\text{err}}, \frac{f}{f_c} \right) = 10 \cdot \log \left[1 + \left(\frac{\alpha_{\text{dB}}}{10^{20}} \right)^2 - 2 \cdot \left(\frac{\alpha_{\text{dB}}}{10^{20}} \right) \cdot \cos \left[2 \pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c} \right) + \theta_{\text{err}} \right] \right] \quad (2.1.5.1)$$

Figure 2.1.5.1 shows the plot of two cancellation performance surfaces as a function of all fundamental performance determining factors for a constant $\lambda_{\text{err}} = 0.3$. The mesh surface corresponds to a cancellation performance as a function of phase error and normalized amplitude frequency deviation for a constant amplitude imbalance of $\alpha_{\text{dB}} = 0.1$ dB. The smooth gradient surface below the mesh surface corresponds to the cancellation performance as a function of phase error and normalized frequency deviation with no amplitude imbalance. Labeled contour lines of the gradient surface are plotted on the XY grid plane of **Figure 2.1.5.1** to help identify the cancellation performance

level for a given phase error and normalized frequency deviation. The surfaces are plotted for a phase error range of $\pm 10^\circ$ and normalized frequency deviation span of 6% bandwidth about the center frequency. The surfaces can be thought of as the 2-D plot of **Figure 2.1.4.3** extruded along the line passing through the center of the XY plane of the phase error (X) and the normalized frequency deviation (Y) axes. The XY line passing through the center of the XY grid in **Figure 2.1.5.1** has a slope of $(1 / 2 \pi \lambda_{\text{err}})$ and is expressed in equation (2.1.5.2) (see **Appendix A5** for derivation).

Figure 2.1.5.1 – Cancellation Performance as a Function of Normalized Frequency Deviation and Phase Error for a Fixed Delay Mismatch ($\lambda_{\text{err}} = 0.3$) and Amplitude Imbalance ($\alpha_{\text{dB}} = 0$ and $\alpha_{\text{dB}} = 0.1$)

$$f_n(\theta_{\text{err}}) = \frac{\theta_{\text{err}}}{2 \cdot \pi \cdot \lambda_{\text{err}}} + 1 \quad (2.1.5.2)$$

The effect of amplitude imbalance on the cancellation performance limits the maximum cancellation level that can be achieved but does not significantly affect the cancellation

bandwidth. The cancellation “V” as a function of normalized frequency deviation is obtained by taking a “YZ-plane slice contour” perpendicular to the phase error axis at a phase error of 0° .

Referring to **Figure 2.1.5.1**, in presence of phase error, the YZ-slice plane could be imagined to be offset from the 0° X-axis reference by the amount of phase error. The cancellation “V” does not change shape but is shifted along the normalized frequency deviation axis at a corresponding offset from the center frequency given by (2.1.5.2). Considering a XZ-slice, would look similar to **Figure 2.1.4.3**, it could be noticed that cancellation nulls re-occur at every multiple of 2π of phase error.

The effect of increasing λ_{err} on cancellation performance has the effect of rotating the surface of **Figure 2.1.5.1** by the central pivot point located at the center of the XY-plane ($\theta_{\text{err}} = 0^\circ$, $f / f_c = 1$). The contour lines get compressed as λ_{err} is increased, meaning the cancellation bandwidth is reduced.

To achieve a cancellation performance better than -30 dB in a 3% bandwidth ($f_n = 1 \pm 0.015$), the amplitude imbalance must be within ($-0.0503 \text{ dB} \leq \alpha_{\text{dB}} \leq 0.05 \text{ dB}$), the phase error must be within $\theta_{\text{err}} = \pm 0.3^\circ$ and the delay mismatch must be better than $\lambda_{\text{err}} = (1/4)$.

2.2 – Feedforward Cancellation System Architecture Description

The feedforward cancellation system (FFCS) architecture evolved into its first implementation driven by the desired functional specification and technological limitations outlined in **Chapter 1**.

Figure 2.2.1 – (a) FFCS Implemented in Base Station RF Radio Front-End, (b) Emissions from PA interfere with the receive signal, (c) PA Noise in receive band is cancelled (d) Single channel in receive band is cancelled.

A potential application of the FFCS is presented in **Figure 2.2.1 (a)**. The FFCS prototype architecture is a self-contained unit that can be placed in an RF path containing an undesired signal portion to be cancelled. The FFCS is designed to be integrated with standard RF communication systems without any major overhaul. The FFCS is attached at the output of the PA and filters the noise in proximity to the transmit signal before being input to the duplexer. The FFCS is used in the same way as a bandpass filter would be used. The FFCS operation is independent of the input signal modulation used, which makes it suitable for being used in state of the art communication system employing modulation schemes resulting in signals with varying envelopes and high peak-to-mean power ratios.

Efficiency and output power of PAs are major factors affecting the cost of RF front-ends. It is critical that the transmit path between the output of the PA and the antenna have

minimal insertion loss. The FFCS is designed to introduce minimal insertion loss at the RF frequencies corresponding to the desired main path signal. The desired main path signal components are represented by the transmit band designated “TX” in **Figure 2.2.1 (b), (c) and (d)**.

The undesired signal components, consisting of output PA noise emissions shown in **Figure 2.2.1 (b)**, overlap with the receive band designated “RX”. In **Figure 2.2.1 (c)**, the noise emissions overlapping with the receive band are cancelled, leaving the signal received by the antenna unmasked.

The FFCS is frequency agile and can be operated as a tunable notch filter. The FFCS can provide a higher cancellation level with reduced bandwidth in the case where the SNR in a single channel in the receive band is to be increased. The cancellation notch center is tuned to the center frequency of the channel to receive, consequently the SNR is optimized for the specific channel in use instead of the entire receive band. The cancellation notch may then be tuned to any other channel as required as illustrated in **Figure 2.2.1 (d)**.

2.2.1 – FFCS Main RF Path

Figure 2.2.1.1 – Feedforward Noise Cancellation System Architecture

The architecture of the FFCS is shown in **Figure 2.2.1.1**. Each signal node used to describe the architecture is labeled. The input RF signal S_{in} arrives at the input port (1). The input signal S_{in} contains both desired spectral components S_i (information signal) and undesired out-of-band spectral components S_n (considered noise and/or interference).

The function of the FFCS is to suppress the undesired spectral components S_n and transmit the desired spectral components S_i at the RF output port (2).

Figure 2.2.1.2 – Main path components and signals

To be a non-intrusive system, the FFCS must have low insertion loss and low distortion characteristics from the RF input port (1) to the RF output port (2). The main path contains only passive components. The passive components in the main path consist of four directional couplers and a delay line. **Figure 2.2.1.2** shows the main path components and the corresponding labels used as nomenclature throughout this thesis. The insertion loss experienced by a signal passing through the main path when the FFCS is in passive mode (i.e. power is OFF) is defined by equation (2.2.1.1) in terms of the S-parameter S_{21} . S_{21} is the transmission coefficient from the input port (1) to the output

port **(2)**. The insertion loss of the main path is dependant on the coupling values of the directional couplers CP1, CP2, CP3 and CP4 and the insertion loss introduced by the main path delay line. The insertion loss of the main path delay line depends on the choice of technology and is a function of the delay line length and frequency. A theoretical approximation of the main path insertion loss calculated with equation **(2.2.1.1)** results in 3 dB for the configuration of the FFCS prototype (see **Appendix B2** for derivation).

$$S21_{\text{main}}(f) := S21_{\text{CP1}} + S21_{\text{CP2}} + S21_{\tau_{\text{main}}}(f) + S21_{\text{CP3}} + S21_{\text{CP4}} + S21_{\text{TXmain}}(f) \quad (2.2.1.1)$$

The transmission characteristic of a directional coupler is a function of the minimum dielectric loss in the coupler and the amount of power that is coupled out of the main path. Equation **(2.2.1.2)** expresses the main path insertion loss characteristics of the directional coupler as a function of dielectric loss **IL** and coupling value **CP**. The logarithmic term in **(2.2.1.2)** is the amount of power that is coupled out of the main path. Equation **(2.2.1.3)** expresses the coupled path insertion loss as a function of insertion loss and coupling value. The coupled path insertion loss is the same whether the signal is being coupled out of the main path as in the case of sampling, or coupled into the main path as in the case of signal injection. A 10 dB coupler with a dielectric loss of 0.25 dB would have a main path insertion loss of 0.78 dB and a coupled port insertion loss of 10.25 dB.

$$S21_{\text{CP}}(\text{CP}, \text{IL}) := \text{IL} + 10 \cdot \log \left(1 - 10^{-\frac{\text{CP}}{10}} \right) \quad (2.2.1.2)$$

$$S31_{\text{CP}}(\text{CP}, \text{IL}) := \text{IL} + \text{CP} \quad (2.2.1.3)$$

The FFCS operates on the principle of destructive interference between a main path signal S_{main} and a cancellation path signal S_{can} . In order to cancel the noise signal S_n , a sample of the input noise signal S_n must be synthesized, phase inverted and re-injected as a cancellation signal S_{can} in the main path. A sample of the noise signal is obtained via a

directional coupler **CP2**. The sampled noise, termed S_{noise} , is then adjusted by the signal controller to match the noise signal components S_n present in the main path and forms S_{can} . S_{can} is re-injected in the main path through the cancellation signal injection coupler **CP3**. The main path signal and cancellation path signal are combined at the point of cancellation in **CP3** resulting in cancellation by vector addition of the two signals.

The delay line τ_{main} assures that the delay experienced by the main signal and the cancellation signal between the point where the noise is sampled and the point where the noise is cancelled are the same. Delay matching between the main path and cancellation path assures the signal fields will overlap in time in the cancellation coupler **CP3** achieving minimum correlation (assuming perfect antiphase of the two signals).

To adjust the cancellation signal such that optimal cancellation occurs requires a measure of the cancellation performance. A pilot tone is used as a reference noise signal to be cancelled and then correlated with the original version of itself. The result of the coherent detection gives enough information about the cancellation process to optimize the cancellation performance of the system as will be explained in **Section 2.2.3**. The pilot tone S'_{pilot} is injected in the noise band through the pilot tone injection coupler **CP1**. The pilot tone undergoes the same processing as the noise signal and is also considered as noise. The signal termed S_x after the cancellation coupler **CP3** is the cancellation process residual. S_x contains both the unmodified desired signal components and some residual noise components which are a function of the cancellation performance. The residual noise components contain the remaining components of the pilot tone. These components are then sampled via the error sampling coupler **CP4** before being input to an I/Q demodulator that performs the coherent detection.

It is important to realize that any signal components introduced in the cancellation path will be re-injected in antiphase in the main path. Consequently, any signal component from the main path leaking into the cancellation path will be suppressed at the point of

Figure 2.2.2.1 – FFCS Cancellation Signal RF Path

Figure 2.2.2.1 show the topology of the signal controller used in the cancellation path of the FFCS. The signal controller contains two main components: two RF Amplifiers and a vector modulator. The RF amplifiers are necessary to overcome the loss associated with the insertion loss in the vector modulator, and the coupling loss of the noise sampling coupler and cancellation signal injection coupler. The gain of the amplifiers are fixed and provide more gain than the losses. To allow the vector modulator to operate at its maximum dynamic range, the resistive π -attenuator provides necessary attenuation to fine-tune the overall gain of the cancellation path. The vector modulator is necessary to modify the noise signal such that it matches its complement in the main path. The vector modulator works by adjusting the I and the Q components of the noise signal, resulting in simultaneous amplitude balance and minimum phase error between the main signal and the cancellation signal, through adaptive adjustments of the I and Q signal amplitude levels.

2.2.2.1 – RF Gain Blocks

The noise is sampled from the main path through the directional coupler **CP2**. The sampled noise level is less than the original noise level traveling in the main path by the coupling factor of the sampling directional coupler plus insertion loss (see equation 2.2.1.3). The vector modulator reduced the sampled noise level by its inherent minimum insertion loss. The cancellation signal injection directional coupler **CP3** further reduces the sampled noise level by its coupling factor. These losses must be compensated for to achieve amplitude balance between the main signal and the cancellation signal at the point of cancellation.

Two gain blocks **G1** and **G2** are inserted before and after the vector modulator respectively. The sum of the two gain blocks overcomes the sum of the losses in the cancellation path, providing acceptable amplitude balance between the main path signal and the cancellation path signal at the point of cancellation.

All active components introduce IMD products and noise in the RF signal path. The IMD products become significant when active devices are operated at high power levels where 3rd order IMD products become significant. To minimize the IMD products, care must be taken to place the components in an order so as to minimize the noise figure of cascade stages. The gain in the cancellation path is distributed before and after the vector modulator such that the active components of the vector modulator are operated in their linear range to minimize the introduction of IMD products.

2.2.2.2 – π Absorptive Attenuator

RF gain blocks are commonly designed to provide fixed gain levels for optimal performance. Tuning the gain of the RF blocks may not be an option. In the case where fixed gain blocks are used, the cancellation path must be designed with a higher gain than necessary. The sum of the gain blocks is more than the sum of the losses in the cancellation path. The gain of the cancellation path must be adjusted to slightly overcome the losses to optimize the dynamic range and reduce the amplitude imbalance and phase error of the vector modulator.

The extra gain in the cancellation path is compensated by a fixed resistive π -attenuator. The resistive π -attenuator uses three resistors layout in π -formation. The π -attenuator provides a constant attenuation and is broadband matched to the characteristic impedance providing zero reflection. The attenuation and matching properties of the π -attenuator shown in **Figure 2.2.2.2.1** are adjusted by choosing the resistors **R1**, **R2** and **R3** appropriately. Details on the design of the π -attenuator are presented in **Appendix C3**.

Figure 2.2.2.2.1 – π -Attenuator

Figure 2.2.2.3.1 – I/Q Vector Modulator Principle of Operation

2.2.2.3 – I/Q Vector Modulator

The I/Q Vector Modulator adjusts the amplitude and phase of the noise signal S''_{noise} by modulating the I and the Q components. The vector modulator has four main components. An input power divider which divides the input signal into two branches, a voltage variable bi-phase reflection type attenuator in each branch and a quadrature hybrid coupler that recombines the signal in each branch in quadrature. The reflection type I/Q vector modulator resembles the typical I/Q modulator composed of mixers in each branch. The only difference between the reflection type I/Q vector modulator is the use of reflection type bi-phase attenuators instead of modulating mixers. This type of

vector modulator is also referred to a vector attenuator since it attenuates vector components of the signal passing through the signal paths [46]. **Figure 2.2.2.3.1** illustrates a typical signal vector component at different stages in the vector modulator. The reflection type topology used for this vector modulator has become very popular in recent years in RF and microwave designs. The control elements used in the voltage variable reflection type modulator is usually PIN diodes or FETs which exhibit variable RF impedance characteristics. The reflection type topology has a higher dynamic range and is better suited for FFCS application as opposed to topologies based on Schottky diodes in a quadrature mixer topology. The reflection type modulator is commonly used in applications ranging from feedforward amplifier linearization, direct RF modulation to signal cancellation systems used in RADARs [47]. Details on each section of the vector modulator used in the FFCS are presented in the next three sections.

2.2.2.3.1 – Input Power Divider

The amplified sampled noise signal S''_{noise} is split in phase and sent along two different paths. S''_{noise} is represented by the polar plot shown in **Figure 2.2.2.3.1 (a)**. One branch is designated the I-branch and is represented by the right path in **Figure 2.2.2.3.1** and the other is designated the Q-branch and is represented by the left path in **Figure 2.2.2.3.1**. The respective signal for each branch coming out of the power divider **CP5** are identified S_{nI} and S_{nQ} and are represented in **Figure 2.2.2.3.1 (b) and (c)** respectively. The signals S_{nI} and S_{nQ} are reduced by 3 dB as a result of the power split. Ignoring the relative delay (phase shift) incurred by the each signal from the input port to each of the output ports of **CP5**, the transmission characteristics for the I and Q signal component are expressed as equation (2.2.2.3.1.1) (assuming negligible insertion loss). The phase shift due to transmission line delay in the power divider can be ignored since it is the same for both signal. Only the relative phase shift difference between the I and Q branch is of importance for the purpose of the analyses performed in this section.

$$S_{21}=S_{31}=\frac{1}{\sqrt{2}} \quad (2.2.2.3.1.1)$$

2.2.2.3.2 – I-Branch and Q-Branch Voltage Variable Bi-Phase Reflection Type Attenuators

Figure 2.2.2.3.2 – Bi-phase Reflection Type Attenuator

The signal S_{ni} is applied to a voltage variable bi-phase reflection type attenuator. The reflection type attenuator modulates the amplitude and the phase of the branch signal S_{ni} . The controlled variable in the reflection type attenuator that makes it suitable for adaptive control, is a voltage variable impedance. The reflection type attenuator configuration is shown in **Figure 2.2.2.3.2**. For simplicity, the coupler is assumed to be lossless, perfectly matched to the system's characteristic impedance at ports (1) and (4) and having infinite isolation.

The attenuator is implemented using a 3 dB quadrature hybrid. The 3 dB 90° hybrid coupler is a symmetrical device. The port notations depend on the configuration of the RF signal connections. To operate the quadrature hybrid as a reflection type attenuator, the RF signal enters at the input port (1) of the quadrature hybrid and the modulated result exits from the isolated port (4). The direct port (3) (0°) and coupled port (2) (-90°) are terminated with identical variable impedances.

The reflection topology illustrated in **Figure 2.2.2.3.2**, uses purely real variable impedances as terminations. A signal incident at the input port is split in two equal

components. One part emerges from the direct port and is delayed by $\lambda/4$ (i.e. 1/4 of the signal wavelength or 90° phase shift) and the other part emerges from the coupled port with no delay (0° phase shift). The voltage wave emerging from the coupled and direct port is reflected by their respective terminating impedances Z_T . The voltage reflection coefficient Γ of the terminating impedances is expressed as a function of the termination impedance in equation (2.2.2.3.2.1). The variable resistors $R1$ and $R2$ are assumed to be equal, hence the reflection coefficients $\Gamma1$ and $\Gamma2$ are the same and denoted Γ .

$$\Gamma(Z_L) := \frac{Z_L - Z_o}{Z_L + Z_o} \quad (2.2.2.3.2.1)$$

With a resistive termination R , the voltage wave emerging from the coupled port and the direct port is absorbed and reflected in proportion to the reflection coefficient presented by the impedance; resulting in insertion loss from the input to the output.

If the terminating impedance is an effective short circuit or an effective open circuit, the voltage wave is reflected completely; resulting in complete power transfer to the isolation port. For any other purely resistive values, a portion of the power will be absorbed by the termination impedance. When the termination impedance is equal to the characteristic impedance of the transmission medium, all the signal power is transferred and dissipated in the termination. No portion of the voltage wave gets reflected.

The voltage wave reflected from the termination is incident on the coupled port and direct port. The voltage waves reflected from the termination re-emerge from the input port and the isolated port. The voltage wave emerging from the input port has zero amplitude since the signal reflected from the direct port is 180° out of phase with the signal reflected from the coupled port when it arrives at the input port, resulting in perfect cancellation. The reflected signal from the coupled port and the direct port arrive at the isolated port in-phase. The transfer characteristics of the reflection type attenuator is

expressed by equation (2.2.2.3.2.2), where R is the variable termination resistor and a 50Ω characteristic impedance is assumed.

$$S_{41}(R) = -1 \cdot \left(\frac{R - 50 \Omega}{R + 50 \Omega} \right) \quad (2.2.2.3.2.2)$$

The ideal characteristics of the reflection type attenuator is shown in **Figure 2.2.2.3.3**, assuming a characteristic impedance of 50Ω . The ideal reflection type attenuator has infinite attenuation dynamic range and allows the perfect relative phase shift of 180° . The operation of the ideal reflection type attenuator is to weight the output signal by a factor ranging from $(+1)$ to (-1) relative to the input signal. The signals S_{nI} and S_{nQ} will be modulated by the respective reflection type attenuator and result in S'_{nI} and S'_{nQ} before being recombined in quadrature to form the complete modulation vector.

Figure 2.2.2.3.3 – Bi-Phase Reflection Type Attenuator Transfer Characteristics

2.2.2.2.3 – 3dB 90° Hybrid Power Combiner

The terminology **I** and **Q** to designate the two signal branches of the vector modulator is used to reflect the method of recombination of the signal in each branch at the output. A 3 dB quadrature hybrid coupler is used for the signal recombination. The I-branch signal S'_{nI} of the vector modulator is incident to the coupled port (0°) hybrid coupler. The Q-branch signal S'_{nQ} of the vector modulator is incident to the direct port (-90°) of the hybrid coupler. The signals incident at both ports are recombined at output port (taken the same as input port (1) in **Figure 2.2.2.3.2**). The signals are added by vector addition which results from two signals being recombined with a -90° relative phase shift between the two. The recombined vector results in the vector sum of S'_{nI} and S'_{nQ} to form the modulated vector S_{vm} . The transmission characteristics of the I-branch signal recombination path and the Q-branch signal recombination path are expressed in equations (2.2.2.2.3.1) and (2.2.2.2.3.2).

$$S_{21} = \frac{1}{\sqrt{2}} \quad (2.2.2.2.3.1)$$

$$S_{31} = \frac{-j}{\sqrt{2}} \quad (2.2.2.2.3.2)$$

The complete vector operation resulting from the independent control of the I and Q vector components makes it possible to position the RF signal vector anywhere in the polar plane. It is hence possible to modify the cancellation path RF signal vector's phase and amplitude to match those of the main path RF signal to be cancelled. The complete transmission characteristic of the I/Q modulator as a function of the I-branch resistive terminations and the Q-branch resistive terminations is expressed in equation (2.2.2.2.3.1) as a function of the variable impedance R_I and R_Q of the I and Q-branch attenuator resistive terminations. S_{41} is the transfer characteristic from (2.2.2.3.2.2) as a function of the respective variable termination.

$$S21_{vm}(R_I, R_Q) = \frac{1}{2} \cdot S41(R_I) + \frac{-j}{2} \cdot S41(R_Q) \quad (2.2.2.2.3.3)$$

Figure 2.2.2.3.4 – I/Q-Plane Vector Modulator Coverage as a Function of Termination Impedance

Equation (2.2.2.2.3.3) can be manipulated to align the I/Q-plane of the vector modulator such that the mathematical results are aligned along conventional Cartesian coordinate system, where the (+I, +Q)-quadrant is in the upper right. Multiplying equation (2.2.2.2.3.3) by the factor (j) rotates the I/Q-plane 90°. The extra rotation is completely arbitrary and ignores the $\lambda/4$ delays in the hybrid couplers. These delays can be ignored inside the vector modulator since both paths are matched in delay. The alignment of the vector modulator's I/Q-plane is chosen such for convenience of analysis. The simplified expression for the I/Q vector modulator is expressed in equation (2.2.2.2.3.4). **Figure**

2.2.2.3.4 shows the I/Q-plane coverage of the I/Q vector modulator and the quadrants relations between impedance values.

$$S_{21\text{ vm}}(R_I, R_Q) = \frac{1}{2} \cdot \left(\frac{R_I - 50\ \Omega}{R_I + 50\ \Omega} \right) + \frac{-j}{2} \cdot \left(\frac{R_Q - 50\ \Omega}{R_Q + 50\ \Omega} \right) \quad (2.2.2.2.3.4)$$

In a practical implementation of this reflection type vector modulator, the minimum theoretical insertion loss is 6 dB for the topology used in the FFCS prototype. The 6 dB minimum insertion loss is due to the 3 dB loss by completely attenuating one branch (pure I or pure Q modulation) and the 3 dB loss in the recombination coupler. The power loss in the recombination hybrid coupler is absorbed by the termination resistor on the isolated port.

2.2.3 – Adaptive Control System

The FFCS is equipped with an adaptive control system (ACS). The ACS is a feedback control loop that senses the current cancellation performance and takes appropriate action to modify the parameters of the vector modulator such that the cancellation performance is optimized. To minimize the complexity of the FFCS architecture, a fully analog control architecture is used.

A few design criteria are considered to determine a suitable control scheme for the FFCS. First, the control system is designed to track slow varying environmental disturbances such as temperature. Since the feedforward cancellation method is a wideband process, under normal operating conditions, the FFCS is tuned to a fixed frequency and is only required to track slow varying environmental changes. The learning time of the adaptive algorithm can be taken as the time required to adjust the control variables such that the target level of cancellation is obtained. The time for a base station to come online may be in the order of a few seconds, from a state of standby. As long as the convergence time is less than the time it takes the base stations systems to come online, the performance is deemed acceptable. The convergence time may be in the order of a few milliseconds to a

few seconds. However, in the case where the filter is required to change channel, the time to adjust to the center frequency step may be a critical design criterion.

The FFCS is not designed to track a wideband signal and cancel the energy of the signal by following the signal modulation. For the FFCS to track a wideband signal modulation would require an adaptation scheme with a learning time as fast as the signal modulation, which could be in the order of 5 MHz to 25 MHz in modern wireless systems. Such performance is prohibited by feedback control mechanism which is essentially a narrowband process, in the order of a few MHz for the application considered. Feedback is limited mainly because of the excessive delay introduced in the feedback loop as the feedback signal bandwidth is increased. Phase shift in the transfer function results in the system becoming unstable. The phenomenon is well stated in a paper by *H. Seidel* [48]: *The inherent limitation of feedback lies within its implicit demand for simultaneity between input and output.* The correction cannot be applied to the input before the output of a process is compared with the input. By the time the output is measured and compared with the input, it is already too late to apply correction to the process. Feedforward cancellation and predistortion technologies are now being used to overcome the very limitations of feedback.

The three main components necessary to control the microwave interference network of the FFCS consist of a sensor to sense the cancellation performance, an actuator to make necessary adjustments to the cancellation signal, and a control loop filter that assures stability and control performance.

Various methods exist to control the variable parameters of the actuator used to adjust the cancellation signal [49]. Two of the most popular approaches consist of cascading an RF attenuator and a phase shifter or using an I/Q vector modulator in the cancellation signal path. An ideal voltage controlled attenuator provides a linear attenuations vs. control voltage with 0° phase shift between the attenuation settings. The ideal voltage controlled phase shifter provides linear phase shift vs. control voltage with 0 dB of attenuation

change between the phase shift settings. In practice, there is phase/amplitude coupling between voltage settings in both components. These undesired coupling effects produce static error in the adaptive control system convergence. Complex control algorithms are required to overcome these effects and require DSP, FPGA or look-up table for implementation. Variable phase-shifter have limited angle tuning range. The phase shift is usually limited to less than 360° or can achieve 360° operation but have a discontinuity in the response or long transition times. Furthermore, the change in phase from 360° to 0° is not continuous, causing a wraparound problem. The phase shifter must return to the 0° state by rotating through -360° . In the implementations where varactor diodes are used, group delay distortion can be introduced as varactors resonate to change their phase. Resistive attenuators experience group delay and phase shift as a function of attenuation change.

The type of vector modulator used in the FFCS architecture performs phase and attenuation change directly with no intermediate steps, reducing the time needed to adjust the vector [7]. The reflection type voltage controlled vector modulator exhibits negligible group delay distortion and absolute delay change and can perform continuous 360° operation without the wraparound problem. PIN diodes exhibit an RF impedance characteristic that is a function of forward bias current. The RF resistance of a typical PIN diode can be varied in the range of a few Ohms to a few kilo-Ohms. The PIN diode is suitable as a control component for use in the vector modulator and forms the actuator of the analog adaptive control system. The I and Q control inputs of the vector modulator can be considered as the weight variables of an adaptive filter. The weight variables are the parameters that must be optimized to achieve optimal cancellation performance.

The cancellation performance is monitored by measuring the residual error signal. The method used to sense the error signal depends on the choice of control algorithm. The control method is kept at a minimum level of complexity by using the least signal processing blocks as possible to generate the control signals that are applied to the vector modulator. The control signals are generated in proportion to the respective I and Q

vector component present in the cancellation residual of the sum of the main signal and the cancellation signal. The control system is adaptive in that the error signal contains the necessary information to drive the weight variables towards optimal cancellation performance.

The surface plot of the cancellation performance defined by the amplitude imbalance and phase error of **Figure 2.1.3.2** can be mapped to a corresponding surface characteristic of the cancellation performance as a function of the I and Q component imbalance between the main signal and the cancellation signal. The I and Q component imbalance between the main signal and the cancellation signal can be expressed as a function of the instantaneous amplitude imbalance α_{dB} and phase error θ_{err} . The expression for the I and Q component imbalance expressed as the normalized difference between the main signal and cancellation signal, are defined in equation (2.2.3.1) and (2.2.3.2) (see **Appendix A6** for derivation).

$$\Delta I(\alpha_{dB}, \theta_{err}) = 1 - \left(10^{\frac{\alpha_{dB}}{20}}\right) \cdot \cos(\theta_{err}) \quad (2.2.3.1)$$

$$\Delta Q(\alpha_{dB}, \theta_{err}) = - \left(10^{\frac{\alpha_{dB}}{20}}\right) \cdot \sin(\theta_{err}) \quad (2.2.3.2)$$

The surface plot of the cancellation performance as a function of the complex error vector is illustrated in **Figure 2.2.3.1**. For clarity, labeled contours of the surface plot are projected on the IQ-plane. The surface plot indicates that the optimal solution to the cancellation problem occurs when the difference in the I and Q components between the cancellation signal and the main signal is minimal. The solution to the cancellation problem consists of minimizing the mean squared error (MSE). Finding the least mean squared (LMS) error of a performance surface governed by variable parameters is well known as the LMS algorithm. The LMS algorithm is one of the main approach used by adaptive systems implemented in the analog domain.

Figure 2.2.3.1 – Surface Plot of Cancellation Performance as a Function of Normalized I and Q Component Difference Between the Main Signal and the Cancellation Signal

The MSE becomes the performance criteria used for adaptation. The MSE performance surface is a N-dimensional paraboloid which has a single minimum, where N is the number of system adaptation variables. In the case of the FFCS, $N=2$, corresponding to the I and Q variables. The LMS algorithm searches the MSE performance surface by updating the parameters of the vector modulator iteratively until it reaches the point where the MSE is minimal (i.e. optimal value) [50]. The convergence of the algorithm is driven in the opposite direction of the instantaneous gradient of the performance surface. The gradient of the performance surface is a N-dimensional vector that points towards the direction of maximum change in MSE and with a magnitude equal to the slope of the tangent of the performance surface.

The error signal must be compared with a reference in order to extract the information that can be used to update parameters of the vector modulator. A training signal is injected in the noise band in the main path before the point where the noise is sampled. The residual of the training signal after the cancellation process can then be detected in a coherent fashion by performing a time correlation of the error signal with the time aligned original reference of this training signal. The result of the correlation is the decomposition of the error signal into an I and a Q component forming the error vector with respect to the training signal. The decomposed error vector serves as a noisy estimate of the performance gradient. Therefore, the decomposed error vector can be directly applied to the LMS algorithm instead of the actual gradient value.

The LMS algorithm is implemented in continuous time by integrating (filtering) the noisy estimate of the performance gradient. An integrator is used to provide an average value for the error vector on a continuous time basis. In the presence of a disturbance, a gradient is established on the performance surface. The negative of the gradient estimate works to push the parameters of the vector modulator towards their new optimal value, resulting in optimization of the cancellation performance. The negative gradient estimate is obtained by negating equations (2.2.3.1) and (2.2.3.2) to obtain (2.2.3.3) and (2.2.3.4) respectively.

$$\varepsilon_I(\alpha_{dB}, \theta_{err}) := 10^{\frac{\alpha_{dB}}{20}} \cdot \cos(\theta_{err}) + 1 \quad (2.2.3.3)$$

$$\varepsilon_Q(\alpha_{dB}, \theta_{err}) := -10^{\frac{\alpha_{dB}}{20}} \cdot \sin(\theta_{err}) \quad (2.2.3.4)$$

The main signal vector V_{main} and cancellation signal vector V_{can} are plotted for a given amplitude imbalance α_{dB} and phase error θ_{err} in **Figure 2.2.3.2 a)**. The normalized error vector is represented by the $(\Delta I, \Delta Q)$ vector in **Figure 2.2.3.2 b)**. The negative of the normalized error vector represents the negative estimate of the performance surface gradient and is shown as the vector ε in **Figure 2.2.3.2 b)**.

Figure 2.2.3.2 – Cancellation Vector Residual I and Q Components (Normalized)

The continuous time integrations used to update the parameter vector $\mathbf{v}_c$ of the vector modulator is expressed in equation (2.2.3.5). In equation (2.2.3.5), the factor μ determines the rate of decrease of the error, hence the rate of adaptation of the control loop. Increasing the adaptation factor μ decreases the convergence time of the algorithm. However, if μ is chosen too large, the algorithm could potentially become unstable. The factor μ also controls the steady-state error of the control loop. μ is effectively the loop gain with respect to a phase locked loop analogy.

$$\mathbf{v}_c(t) = \mu \cdot \int_{-\infty}^t \epsilon(\alpha_{dB}, \theta_{err}, u) du \quad (2.2.3.5)$$

Where: $\epsilon = \epsilon_I + j \cdot \epsilon_Q$

The resulting control vector $\mathbf{v}_c$ of equation (2.2.3.5) is composed of the I component weight variable $\mathbf{v}_{cI}$ and the Q component weight variable $\mathbf{v}_{cQ}$. The control vector components $\mathbf{v}_{cI}$ and $\mathbf{v}_{cQ}$ represent the control voltage applied to the vector modulator. The control vector is the current best weight factor estimate to produce the optimal cancellation signal.

The FFCS has two identical control loops which operate independently to optimize the weight variables for the I and Q components. One channel is designated the I-channel and the other is designated the Q-channel. Only the I-channel control loop is analyzed in this section. The same theory applies to the Q-channel.

Figure 2.2.3.3 – FFCS Adaptive Control System Block Diagram

The process flow for the adaptive control system is summarized in **Figure 2.2.3.3**. Referring to **Figure 2.2.3.3**, a training signal $R(s)$ is input to the system. The training signal is injected in the same frequency band occupied by the noise to be cancelled. The training signal serves as a reference signal to align the I/Q plane of the vector modulator and the vector demodulator. The training signal is also the basis used to generate the error signal from which the control information can be extracted. The training signal is applied to the microwave interference network which is the process being controlled. One branch (cancellation path) of the process modifies the signal $R(s)$ in phase, amplitude and delay. The cancellation process is affected by an unknown disturbance characteristic which depends on the frequency of the training signal. In the model of **Figure 2.2.3.3**, the disturbance may be introduced additively or as a product factor as $P(s)$. The phase and amplitude variations can also be considered as variations in the complex components I and Q of the main signal with respect to the cancellation signal. The vector modulator is represented by the signal control actuator $M(s)$ in the cancellation path. The vector modulator characteristic reflects the state of the two weight variables that consist the control vector (I and Q components control voltages).

The summation of the two branch signals $\mathbf{R(s)}*\mathbf{P(s)}$ and $\mathbf{R(s)}*\mathbf{M(s)}$ result in the cancellation signal residual $\mathbf{E(s)}$. An error signal is detected by a vector demodulator with a detector characteristic represented by $\mathbf{D(s)}$. The training signal $\mathbf{R(s)}$ applied to the I/Q vector demodulator is used as the reference plane to decompose the error signal $\mathbf{E(s)}$ into an I and a Q component. The error signal $\mathbf{E(s)}*\mathbf{D(s)}$ is applied to the loop filter $\mathbf{F(s)}$. The loop filter acts to filter out the noisy negative gradient estimate and essentially hold the value of the current best estimate of the weight variables. The output of the loop filter is then applied to the vector modulator. Stable loop operation drives the error signal to a minimum determined by the final value of the error transfer function of the loop.

2.2.4 – Adaptive Feedback Control Loop Topology

The information necessary to optimize the cancellation performance is obtained from the baseband outputs of the I/Q demodulator. The presence of a signal at the baseband output of the I/Q demodulator indicates that cancellation performance is not optimal.

Referring back to **Figure 2.2.1.1**, the pilot tone signal $\mathbf{S_{pilot}}$ is input to the system where it is split by the in-phase power divider **CP9**. The split signals $\mathbf{S'_{pilot}}$ and $\mathbf{S''_{pilot}}$ are the same. $\mathbf{S'_{pilot}}$ is injected in the main path through coupler **CP1** and $\mathbf{S''_{pilot}}$ goes through the delay line $\mathbf{\tau_{ref}}$ before being input to the LO port of the I/Q demodulator.

The pilot tone component injected in the main path is placed somewhere in the noise band to be cancelled. The pilot tone is the target component that the ACS of the FFCS is driven to cancel. Therefore, the pilot tone component goes through the same process variations as experienced by the noise components around the pilot tone frequency. The pilot tone undergoes a series of modifications. Modifications are impressed by unknown or uncontrolled disturbances that may be caused by environmental disturbances or frequency dependent characteristics of the RF paths. The pilot tone proceeds to the point of cancellation. The main path component is summed with the cancellation path

component. The pilot tone residual S_{error} is attenuated by the coupling factor of the error sampling coupler **CP4** and is applied to the RF port of the I/Q demodulator.

The I/Q demodulator decomposes the error signal into baseband I and Q error gradient signal estimates using the pilot tone reference signal applied to LO port of the I/Q demodulator. The reference pilot tone is essential to align the I/Q plane of the modulator and the demodulator. The pilot tone reference driving the LO of the I/Q demodulator enables extraction of the I and Q information from the error signal in a coherent fashion. It is also assumed that the I/Q plane of the I/Q modulator is aligned with the I/Q plane of the I/Q demodulator by using the pilot tone reference signal as a calibration standard (refer to **Section 3.3**).

When the LO port drive signal S_{ref} and the RF port drive signal S_{error} of the I/Q demodulator are at the same frequency, the baseband output characteristic of the I/Q demodulator can be expressed by equations (2.2.4.1) and (2.2.4.2). The baseband output characteristic v_{eI} and v_{eQ} of the I/Q demodulator is expressed as a function of the error signal power S_{error} (dBm with reference to 50Ω characteristic impedance) and error signal phase θ_e with respect to the pilot tone S_{ref} . K_D is the conversion gain factor of the I/Q demodulator obtained from the hardware model at constant LO drive level. K_D is expressed in units of dB. The baseband output of the I/Q demodulator has a sinusoidal characteristic. The amplitude of the baseband output is governed by the gain factor K_D and the power of the error signal S_{error} . The power of the amplified error signal is hence converted to the instantaneous voltage equivalent in a 50 Ω system.

$$v_{eI}(\theta_e, S_{\text{error}}) := \sqrt{\left(\frac{S_{\text{error}} + K_D}{10^{10}} \text{ mW} \right) \cdot (50 \Omega)} \cdot \sqrt{2} \cdot \cos(\theta_e) \quad (2.2.4.1)$$

$$v_{eQ}(\theta_e, S_{\text{error}}) := \sqrt{\left(\frac{S_{\text{error}} + K_D}{10^{10}} \text{ mW} \right) \cdot (50 \Omega)} \cdot \sqrt{2} \cdot \sin(\theta_e) \quad (2.2.4.2)$$

Figure 2.2.4.1 is a plot of typical baseband output characteristic of the ideal I/Q demodulator as a function of phase difference when the two signals compared are single tones at the same frequency as is the case for the FFCS. In **Figure 2.2.4.1**, the detector gain K_D is 40 dB and S_{error} is at -40 dBm. The characteristic of Figure 2.2.4.1 closely resembles the actual hardware characteristic of the I/Q demodulator.

Figure 2.2.4.1 – Ideal I/Q Demodulator Baseband Output Characteristics as a Function of Phase Difference Between S_{error} and S_{ref}

In the case where S_{error} is in-phase with S_{ref} , the I component baseband output is at a maximum and proportional to the power level of the error signal. When the two signals are in quadrature, the Q component is at a maximum voltage level and proportional to the power level of the error signal. Taking S_{ref} as the reference (0°), when the phase rotation of S_{error} (i.e. increasing θ_e) is clockwise on the I/Q plane then the I/Q demodulator output is counter-clockwise.

The demodulated I and Q components provide enough information to characterize the cancellation performance. In the ideal case where the error signal is driven to $-\infty$ dBm, the baseband outputs will also be 0 V indicating maximum cancellation performance has been achieved. The results provided by the I/Q demodulator is a direct indication of which component of the I/Q vector modulator should be tuned and whether it should be increased or decreased.

It is important to understand the vector modulator and I/Q demodulator transfer characteristic to interpret the cancellation error signal information in the FFCS. The transfer characteristic from the I/Q vector modulator to the I/Q demodulator is obtained by coupling the output of the vector modulator directly to the RF input of the I/Q demodulator and applying the same test signal through the vector modulator and the LO input of the I/Q demodulator. In the context of the FFCS, the test setup is achieved by removing the main path delay line from the circuit.

Figure 2.2.4.2 illustrates the transfer characteristic for a single tone test signal. Assuming a one-to-one correspondence between the control input of the vector modulator and detector outputs, a given vector modulator control input (v_{cI} , v_{cQ}) in **Figure 2.2.4.2 (a)** results in the I/Q demodulator baseband output (v_{eI} , v_{eQ}) shown in **Figure 2.2.4.2 (b)**. The I/Q reference plane of the FFCS is set with the reference signal driven into the LO port of the I/Q demodulator. The I/Q reference plane is therefore relative to the instantaneous value of the pilot tone reference signal incident at the LO port of the I/Q demodulator. **Figure 2.2.4.2** shows that *increasing* the I-component control voltage of the vector modulator *decreases* the I component of the modulated signal and the I/Q demodulator sees a *decrease* in the detected I component. *Increasing* the Q-component control voltage of the vector modulator results in an *increase* in the Q component of the signal being modulated while the I/Q demodulator sees a *decrease* in the detected Q component. Therefore, it can be said that the I component transfer characteristic of the vector modulator to I/Q demodulator are the same while the transfer characteristic of the Q component is inverted. A vector rotated counter-clockwise by the vector modulator control inputs would appear to rotate clockwise at the I/Q demodulator baseband outputs.

Hence, it can be realized that the characteristic of **Figure 2.2.4.2**, which represents the output of the I/Q demodulator, is essentially the noisy estimate of the error gradient formed by the multiplication of the error signal with the reference signal. Therefore, the output of the I/Q demodulator can be used directly to drive an integrator to produce the

control signal for the vector modulator control input to reduce the I and Q components of the signal being modulated.

Figure 2.2.4.2 – Vector Modulator Input to I/Q Demodulator Output Transfer Characteristic

Figure 2.2.4.3 – Sample Cancellation Scenario

In a typical cancellation scenario illustrated in **Figure 2.2.4.3**, the signal S_{main} is summed to the signal S_{can} . Prior to summation, S_{can} is modulated by the vector modulator. The signal S_{main} can be a vector arbitrarily placed in the I/Q-plane. The signal S_{error} that results from the summation of S_{main} and S_{can} is shown in **Figure 2.2.4.3 b)** along with the error components v_{eI} and v_{eQ} , as seen by the I/Q demodulator. In the next time step, the negative error gradient ϵ_I (modified by some proportionality factor and cumulated) is fed-back and applied to the control input of the vector modulator. The I component of S_{can} then undergoes an increase from its current state. In the same respect, the positive error gradient ϵ_Q contributes to increase the Q component of S_{can} from its current state. The

final result of a series of such iterations produces a successful cancellation process and the optimal S_{can} becomes the same amplitude but opposite phase as S_{main} . The weight variables v_{cI} and v_{cQ} are optimized when the power of S_{error} is minimized or when time average product of S_{error} and S_{ref} is null. In both cases, the baseband output error estimates will have an average of zero.

Because in practice, the correspondence between the control input of the vector modulator to the I/Q demodulator output is not one-to-one, the baseband output of the I/Q demodulator (negative gradient estimate ϵ) must be integrated to determine the optimal control signal applied to the vector modulator. The baseband processor performs two functions. It amplifies the error signal and integrates the negative error gradient to produce the vector modulator control signal. The control system contains one integrator and is therefore a first order feedback loop. The general system level topology of the baseband processor is shown in **Figure 2.2.4.4**.

Figure 2.2.4.4 – Baseband Processor System Block Diagram

The baseband processor consists of the three main function blocks illustrated in **Figure 2.2.4.4**. The baseband output v_e from the I/Q demodulator is amplified through the baseband amplifier **K1**. The amplified error signal v_s is then input to the loop filter **F(s)**. **F(s)** is an integrator that filters the noisy gradient estimate. The output of the integrator v_{cp} is buffered through the gain stage **K2** to produce the control signal v_c .

The cancellation performance limit that can be achieved by the FFCS is governed by the loop gain which essentially sets the order of magnitude of the steady state error. The loop bandwidth is governed by the loop gain and the location of the pole set by the time constant of the integrator. Any noise present at the control input of the vector modulator

will modulate the I and Q components of the cancellation signal and degrade the similarity between the main signal and the cancellation signal at the point of cancellation.

The control loop parameters set the dynamic performance of the FFCS in terms of noise figure of the modulator, lock time, steady state error and dynamic error. The lock time of the FFCS is essentially the time it takes the adaptive control system to learn the optimal control values. The steady state error that results in the baseband section impacts the cancellation performance. The dynamic error impacts the instantaneous cancellation performance during a sweep of the pilot tone.

The characteristic of the I/Q vector modulator is highly non-linear due to the RF impedance characteristic which is a non-linear function of bias current. The dynamic performance of the control loop of the FFCS is determined by simulation using a simplified linear model of the control loop.

2.3 – FFCS Architecture Simulation Using Agilent's ADS

Prior to hardware implementation, system level simulations of the FFCS were performed using Agilent EEsof EDA's Advanced Design System (ADS) to validate the design architecture. ADS was chosen for its robust simulation engine, ease of use and complete design flow project management capabilities. The various ADS simulation engines allow the freedom to simulate circuits from ideal system models, circuits with ideal and mixed hardware models, to complete hardware simulation and layout vs. schematic simulation.

ADS has a few different simulation controllers that are used to analyze the circuit. The choice of simulation depends on the performance characteristics to be determined. The Harmonic Balance simulation controller uses nonlinear harmonic-balance techniques to find the steady-state solution in the frequency domain. Harmonic balance is typically used for mixer, oscillator, power amplifier, and transceiver designs. The Transient/Convolution simulation controller solves a nonlinear circuit entirely in the time

domain using simplified models to account for the frequency-dependent behavior of distributed elements. It is typically used for mixer, power amplifier, and switching circuit designs. The S-Parameter simulation controller provides linear S-parameter, linear noise parameters, transimpedance, and transadmittance simulation. The S-parameter simulation is typically used for filter, oscillator, and amplifier designs.

A significant portion of the FFCS simulation require use of the Circuit Envelope simulator. The Circuit Envelope simulation controller uses a combination of frequency-domain and time-domain analysis techniques to yield a fast and complete analysis of complex signals such as digitally modulated RF signals. It is typically used for mixer, oscillator, power amplifier, and transceiver, and phase-locked loop designs. Details on the envelop simulator can be found in the ADS 2002 product documentation, specifically in the simulation manual.

In a circuit such as a PLL, the time response of the controller is much larger than the period of the carrier. To simulate a PLL in operation, a typical transient analysis would require calculating the state variables of the circuit at time steps in the order of fraction of the carrier to yield a solution. In a system operating at 1 GHz with control time response in the order of 1 ms, simulated with a reasonable time step of 0.1 ns, would require the system state variable to be calculated 100 million times. This requires tremendous computation capabilities to yield analysis in decent time. The control system of the FFCS operates at baseband (DC to a few MHz) while the high frequency RF systems operate upwards of 800MHz. The important information to extract from the simulation is the RF signals amplitude and phase at various points in the circuit as the adaptive circuit converges towards optimal cancellation. The envelope simulator is the best simulation tool to yield this information. The FFCS can therefore be simulated with very good accuracy in the order of a few seconds to less than five minutes on a SUN Ultra 10.

2.3.1 – FFCS Architecture System Test Bench

harmonic simulator is activated, the two DC sources are activated and the baseband components are deactivated to open the control loop. This ensures that the feedback loop is not trying to adjust the voltage nodes labeled **V_cI** and **V_cQ** that are required to be fixed at a constant DC levels.

Some important variable definitions are declared in the lower left box of **Figure 2.3.1.1**. Variables declarations allow quick reference of values during CAD and avoid mistakes that could be hard to find throughout the hierarchy of sub-circuits. Under the variable component **VAR1**, the pilot tone frequency that is applied to the pilot tone input port is declared.

The substrate information of the material used for the PCB hardware implementation is detailed in the substrate information component **MSub1**. The line width required for RF transmission lines having characteristic impedance of 50Ω at the design center frequency of 836.5 MHz was obtained with ADS's LineCalc tool. LineCalc is a tool used to calculate the RF transmission line parameters given the substrate information. Under the variable component **VAR2**, the calculated line width is **LineW**. **LineW** and **MSub1** are used to model actual 50Ω microstrip RF transmission lines using transmission line components labeled **MLIN** in the schematic. The variable **LineL90** is the length of transmission line through which a 90° phase shift at the center frequency of 836.5 MHz results. Having the wavelength accessible through a variable is convenient to simulate the effect of delay mismatch as a function of fractions of wavelength.

Other variables are declared under **VAR2**. **VcI** and **VcQ** are used to hold the values of the control voltages applied to the vector modulator once the cancellation performance is optimized with the envelope simulation. These variables set the node voltages of **V_cI** and **V_cQ** by means of the two DC sources, located in the simulation controllers section, for use during a harmonic balance simulation. **LineL90** is the length of transmission line that results in a relative 90° propagation phase shift of the 836.5 MHz frequency components.

The variables x and y are multipliers used to set the length of transmission line necessary for matching the pilot tone reference path and main path delays respectively. These values were obtained using the ADS optimization feature. ADS optimization simulation controller component is useful to find the optimal design parameter of a given component. In this case, the optimizer was used in conjunction with the gradient min/max algorithm to find the optimal delay line length to match the delay from the pilot tone input to RF port of the I/Q demodulator through the cancellation path with the main path delay line deactivated. The optimizer was also used to match the delay through the cancellation path and the main path. The procedure devised to design the actual delay lines is elaborated in **Section 3.3**, and was performed in the same fashion using the ADS optimization controller and using the same design goals.

The reader should take note that the test bench is an approximate model to the actual hardware implementation. The system model was developed to test the main functional behavior of the system and to gain better insight into specific areas of the design affecting the cancellation performance. For example, not all the RF transmission lines are included in the system schematic. This explains why the pilot tone reference path delay line multiplier x is smaller than the main path delay line multiplier y in the simulation test bench. The FFCS system model test bench was not designed to include layout specific information.

2.3.1.1 – Component Hardware Models Description

Most system components used in the ADS simulation are ideal models. The control components however, were modeled to behave similar to the actual hardware evaluation of the respective component or subcircuit used in the implementation. The simulation results better reflect the actual physical phenomenon. Brief details on the design of component models adapted to the hardware behaviors are worth mentioning to better

understand the simulation results. The models reviewed are: the I/Q demodulator, the vector modulator and the baseband processor.

Figure 2.3.1.1.1 – Vector Modulator

The vector modulator uses the same design as explained in **Section 2.2.2**. The model used is illustrated in **Figure 2.3.1.1.1**. The extra transmission line **TL1** introduces a sample delay in the vector modulator model and may be set to any length. The delay match between the main path delay line and the reference path delay line is influenced by **TL1** delay. The couplers used in the hardware implementation exhibit variable amplitude variation as a function of frequency. The two-port frequency defined device (FDD2P) labeled “HW_Amplitude_Imbalance” is used to introduce amplitude variation as a function of frequency to approximate the non-ideal wideband amplitude variation inherent in the directional couplers. The FDD in ADS is used to design equation-based, user-defined non-linear components. Introducing the amplitude variance as a function of frequency in the simulation produces results that are closer to the actual hardware measurements.

The voltage variable reflection type vector attenuator model for **VVA_I** and **VVA_Q** used in the vector modulator’s I and Q branch respectively, is illustrated in **Figure**

2.3.1.1.2. S-parameter data supplied by the manufacturer of the 90° quadrature hybrid is used to model the coupler. Two coupling capacitors are used on the 0° and 90° ports of the hybrid coupler to isolate the RF paths from the DC bias voltage on the PIN diodes. Although actual PIN diode models are not used in the vector modulator model, the effect of these capacitors must be included in the model to reflect the hardware behavior of the reflection type attenuator. The terminating impedance on the 0° and 90° ports of the hybrid coupler is not purely real in the presence of the coupling capacitors.

Figure 2.3.1.1.2 – Voltage Variable Reflection Type Attenuator

No accurate PIN diode model could be used in the ADS simulation. It was sufficient for the purpose of analysis to develop an approximate non-linear model of the behavior of the RF impedance of a PIN diode in response to a forward DC bias current (or voltage). The voltage variable impedance is modeled using an ADS frequency domain device (FDD) component. The FDD used to model a voltage variable impedance is a 3-port device that describes current and voltage spectral values on port 2 and 3 in terms of algebraic relationships of voltage and current spectral values on port 1. Envelope and harmonic balance simulations can then be performed with the behavioral model defined with an FDD component. The RF impedance model essentially establishes an impedance to the spectral components at **Pilot_Tone_Freq** and **RF_Freq** frequencies on ports 2 and 3, proportional to the DC (baseband) voltage. The PIN model exhibits an approximate 22

Ω DC impedance due to the parallel combination of the 43.2Ω diode bias resistors.

The presence of the second RF frequency component **RF_Freq** is to allow testing the system with two fundamental tones. If the FFCS was to be simulated with more than two fundamental tones, the model designed shown in **Figure 2.3.1.1.3** would have to be adapted to include the extra tones.

Figure 2.3.1.1.3 – Voltage Variable RF Impedance Model

Figure 2.3.1.1.4 – I/Q Demodulator

The I/Q demodulator model is shown in **Figure 2.3.1.1.4**. The design models the behavior of the I/Q demodulator used in the hardware implementation. The RF signal is

amplified by an ideal RF gain block that provides 20 dB of gain. The amplified RF signal is then split into two in-phase components. Each branch signal is applied to the RF port of an ideal mixer. The LO input signal is split into an in-phase and a quadrature component and applied to the LO port of the ideal mixers to form the I and the Q branch. The upper sideband mixing products are removed by low-pass filtering using a Butterworth filter with a 3-dB bandwidth of 20 MHz. Only the baseband portion of the mixer output in each branch is kept. The filtered output of each branch is amplified by ideal gain blocks that have a gain of 22.9 dB. The output impedance of the gain blocks are referenced to $50\ \Omega$ therefore, terminating the output of the amplifiers with a $50\ \Omega$ impedance to assure proper gain conversion. The input of the following baseband processor is high impedance to avoid loading the demodulator model.

Figure 2.3.1.1.5 – Ideal Mixer Model

The ideal mixer model used in the I/Q demodulator is also defined using an FDD component and is shown in **Figure 2.3.1.1.5**. The mixer is designed to produce the correct baseband output corresponding to an RF input and LO input at the same frequency. The mixer performs frequency multiplication of the two vectors.

In the baseband processor shown in **Figure 2.3.1.1.6**, two ideal opamp gain stages are used to approximate the behavior of the baseband amplifier followed by the charge pump buffer amplifier. The input amplification stages have a combined voltage gain of 85 dB. A limiter component is used to limit the voltage at the output of the ideal opamp stages to $\pm 4\text{V}$ which closely represents the actual output swing of the charge pump driver.

Figure 2.3.1.1.6 – Baseband Processor Model

The charge pump is modeled from the actual circuit implementation and uses the transistor Spice models supplied by the manufacturer. Details on the charge pump design are presented in **Section 3.1.6.3**. The output stage of the baseband processor is an opamp circuit with a voltage gain of 1.2. The opamp circuit acts as a buffer to drive the low impedance PIN diodes from the high impedance loop filter output. More details on the PIN diode driver is presented in **Section 3.1.6.4**.

The **ResetSwitch** components are used to set the initial conditions to zero (0 V) on the nodes containing capacitors during simulation. If the **ResetSwitch** is not present, the initial conditions of the simulation are undefined and may be non-physical.

2.3.2 – Envelope and Harmonic Balance Simulation Results

The RF performance at the pilot tone frequency as well as the analog baseband transient characteristics are obtained using the envelope simulator. The simulations were used to validate the FFCS concept, evaluate the performance and investigate the behaviour before the actual hardware implementation. The simulation results establish a bench mark used to compare with hardware implementation.

The envelope simulation was setup to run at the design center frequency of 836.5 MHz, with 1 microsecond time step intervals for a total simulation time of 1 millisecond. The results of the envelope simulation are shown in **Figure 2.3.2.1**. The initial state of all node voltages are set at 0 V. The simulation is analogous to the FFCS being powered on from a state of zero energy. The simulation may also represent the response to a disturbance step in the operating conditions at $t=0 \mu\text{s}$ caused by a step in the pilot tone.

In **Figure 2.3.2.1 (a)**, the cancellation performance level takes $120 \mu\text{s}$ to converge to the optimal value of -94 dB . The actual amplitude balance and phase error as the FFCS adapts is shown in **Figure 2.3.2.1 (b)**. The polar plot in **Figure 2.3.2.1 (c)** shows the vector behavior as the FFCS progresses towards convergence. The circle in the lower right quadrant represents S_{main} . The upper right curve represents S_{can} . S_{can} ultimately converges towards the negative value (anti-phase) of the S_{main} . The curve converging to the center of the plot is the cancellation residual signal S_x .

The steady state values at the last time step are shown below plot **Figure 2.3.2.1 (b)** and are calculated using the equation below the polar plot of **Figure 2.3.2.1 (c)**. The vector voltages were measured before they entered the cancellation coupler. The equations calculate the compensation factors needed to represent the vectors by including the effect of the cancellation coupler to represent the cancellation performance information accurately. **CancelFact** is used to compensate S_{main} for the main path insertion loss through the cancellation coupler. **CancCPLFact** is used to compensate S_{can} and is the

coupling factor of the coupler. The vector S_{can} is multiplied by the complex operator i to account for the 90° phase shift through the cancellation coupler's coupled port into the main path.

The RF performance obtained from the envelope simulations shows the cancellation performance with time. The vectors shown in the polar plot illustrate the adaptive nature of the FFCS. The cancellation signal aligns 180° deg out of phase with the main signal. Only the signal component present at the pilot tone frequency in this simulation.

Figure 2.3.2.1 – FFCS RF Performance Results from ADS Envelope Simulation

Shown in **Figure 2.3.2.2**, the baseband performance is also obtained through the envelope simulation. The I-channel baseband signals are represented by a solid curve and the Q-channel signals are represented by a dot-dash curve. **Figure 2.3.2.2 (a)** illustrates the error voltage present at the output of I/Q demodulator. The error

components V_{eI} and V_{eQ} are reduced to the steady state error governed by the loop gain.

The output of the baseband amplifier stages is the control voltage input to the charge pump shown in **Figure 2.3.2.2 (b)**. The input to the charge pump V_{c_cp} is limited by the voltage limiter which is used to model supply rail limiting that occurs in actual hardware implementation. The charge pump input control voltage of the I and Q-channel baseband processor is saturated until the error signal at the input of the baseband amplifier falls within the linear region of the amplifier stages and no longer drives the charge pump driver amplifier into saturation. The buffered output of the charge pump is shown in **Figure 2.3.2.2 (c)**. The output of the PIN diode driver stage of each channel is the control signal V_{cI} and V_{cQ} applied to the reflection type attenuators in the vector modulator. The steady state values of FFCS baseband processor signals are shown at the right of each plot in **Figure 2.3.2.2**.

Figure 2.3.2.2 – FFCS RF Performance Results from ADS Envelope Simulation

Figure 2.3.2.3 shows the wideband performance obtained from the harmonic balance simulation at the FFCS control state obtained from the previous envelope simulation. The cancellation performance V-shaped characteristic as a function of frequency is shown in **Figure 2.3.2.3 (a)**. The amplitude imbalance in (dB) and phase error in (degrees) is shown in **Figure 2.3.2.3 (b)** as a function of frequency. **Figure 2.3.2.3 (b)** is a very good representation of the amplitude imbalance and phase error disturbances that are introduced if the pilot frequency is stepped at an offset of the center frequency. The deviation in amplitude imbalance and phase error from the current point of operation resulting from a step in the pilot frequency would represent the frequency dependant disturbance $P(s)$ discussed at the end of **Section 2.2.3**.

Wideband performance results are obtained from the Harmonic Balance Simulation with VcI and VcQ fixed to the converged values with the pilot tone centered at 836.5 MHz. The converged components are the final values obtained from the Envelope Simulation.

Figure 2.3.2.3 – FFCS RF Performance Results from ADS Envelope Simulation

2.3.3 – Adaptive Control System Linear Model Performance

The performance and characterization the adaptive control system is obtained with a simplified system analysis of the baseband processor. Open loop AC analysis and closed loop response to a step, ramp and sinusoidal input disturbance are obtained in simulation to evaluate the performance of the control system. The analysis of the control architecture demonstrates transient response, loop bandwidth, steady state error and stability.

Figure 2.3.3.1 – Adaptive Control System Simulation Test Bench

The simulation test bench used to evaluate the performance is shown in **Figure 2.3.3.1**.

To characterize the performance of the control system, the loop is linearized by only considering the characteristic of the baseband processor and closing the loop. It is assumed that a small change in disturbance invokes a small change of the control signal. The operating point is assumed such that the combined sensitivity of the vector modulator and I/Q demodulator can be taken as unity. The baseband processor has a non-inverting characteristic. The negative feedback loop is closed by multiplying the output of the baseband processor by -1 using the **VMult** component labeled “INVERT”.

2.3.3.1 – Open Loop AC Analysis of Baseband Processor

Figure 2.3.3.2 – Open Loop AC Analysis of Baseband Processor

The open loop AC response of the baseband processor is obtained using the AC simulation component. An AC analysis is performed from 1 MHz to 20 MHz. The results are shown in **Figure 2.3.3.2**.

In **Figure 2.3.3.2**, the AC analysis shows that the open loop gain is 104.58 dB. The open loop bandwidth is close to 200 kHz with a 3 dB bandwidth of approximately 3 Hz. The high gain results from the high gain chosen for baseband amplifier to assure sufficient error signal amplification to fully switch the charge pump. The gain can be broken down into 85 dB coming from the input amplifier stages, 20 dB loss in the voltage limiter, 38 dB gain in the charge pump and 1.67 dB from the PIN diode buffer. The gain of the actual hardware implementation is limited to a more realistic value of 60 dB. The phase margin of 66.45° at unity gain assures that the loop is stable. Since the control system uses a first order loop, the adaptive control system of the FFCS is unconditionally stable.

2.3.3.2 – Close Loop Transient Response of Feedback Control Loop

The loop response to a step, ramp and sinusoid disturbance input. **Figure 2.3.3.2.1** shows the response to a 1 V step disturbance. The disturbance signal v_d , shown in **Figure 2.3.3.2.1 a)**, is generate using the **VtStep** source component labeled “S_Step”, with the other source components de-activated. The step is generated after an initial delay of 100 μ s and has a rise time of 1 ns, which is much smaller than the time constant of the loop filter.

The step response shows that the loop filter has a time constant of approximately 310 μ s. The steady state error is 11.43 μ V which is 99 dB down from the disturbance step and close to the maximum level of cancellation achieved (refer to **Figure 2.3.2.1**). The response shows that the control signal **Vc_cp** applied to the charge pump is saturated until the error signal is reduced to the steady state error. The steady state error of the control loop sets the maximum cancellation level that can be achieved at a fixed pilot tone frequency.

The loop response to a ramp input shown in **Figure 2.3.3.2.2**, is obtained by setting a large rise time to the **VtStep** component. The ramp signal has a slope of 1000 V/s. The loop response to the ramp disturbance shows a steady state error of 644 μV in tracking performance. The steady state error to a ramp input indicates that the tracking performance of the control system would limit the cancellation performance under swept pilot tone conditions. In the case of the ramp input, the cancellation performance of **Figure 2.3.3.2.2** would be reduced by 35 dB when compared with that of **Figure 2.3.3.2.1**.

Figure 2.3.3.2.1 – Transient Simulation: Response to Step Disturbance

The response to a sinusoid disturbance is shown in **Figure 2.3.3.2.3**. The sinusoid response shows the control loop response for a case where the pilot tone control signal would be modulated. The 1 V sinusoid has a frequency of 100 Hz. The error does not have a steady state value and varies with the disturbance. The error is less near the peaks of the sinusoid since the integrator has time to “catch up” with the disturbance signal during the transition from a positive slope to a negative slope. The error is highest where the rate of change of the disturbance signal is greatest. The slope of the sinusoid as it crosses 0 V is higher than the rate of change of the loop filter, hence the integrator cannot keep up. The tracking error for a sinusoid input can be reduced by increasing the loop bandwidth.

Figure 2.3.3.2.2 – Transient Simulation: Response to a Ramp Disturbance

Figure 2.3.3.2.3 – Transient Simulation: Response to a Sinusoid Disturbance

2.4 – Cancellation Performance Practical Limitations

In practice, it is impossible to have perfect control over component and subcircuit parameters. Factors other than amplitude imbalance, phase error and delay mismatch can potentially limit the cancellation performance of the FFCS. This section describes some limiting factors that arise because of practical implementation issues. The limited modulation capabilities of the vector modulator may limit the maximum cancellation level achievable in a particular condition where the weight factors are required to

modulate near the pure I or pure Q axes. The importance of matching the reference path delay line is also overviewed.

2.4.1 – Limited I and Q Component Dynamic Range in Vector Modulator

When the vector modulator characteristics are specified, it is important to take into account the minimum insertion loss and the maximum insertion loss in each of the I and Q branches. The dynamic range of the reflection type attenuator directly impacts the range of possible vector modulations achievable. The attenuation dynamic range of the voltage variable reflection type attenuator is limited in part by the RF impedance limits used as terminations on the coupled and direct ports of the 3 dB quadrature hybrid couplers and in part by the isolation of the hybrid coupler.

As was shown in **section 2.2.2.3**, the smallest insertion loss state of the reflection type attenuator is limited by the smallest and largest value of the termination impedance. The extreme limits of termination impedance are desirable to achieve the ideal short or open circuit condition at the terminated ports of the coupler. The prototype developed uses small signal PIN diodes as voltage variable impedances. In practice, the PIN diode has a finite range of possible RF impedance values it can achieve within its rated operable forward bias operating conditions. PIN diodes cannot produce perfect short circuits or open circuits. The RF impedance characteristics of the PIN diodes used in the FFCS prototype can be approximated by the equation (2.4.1.1.1), plotted in **Figure 2.4.1.1.1**. The impedance of the small signal PIN diodes used is a function of forward bias current and can be varied from approximately 2 Ω to 1.6 k Ω .

$$R_{\text{pin}(i)} := 1.5 \, \Omega + \left(\frac{12}{1^{0.9}} \right) \Omega \quad (2.4.1.1.1)$$

The upper and lower limit of termination impedance set a minimum insertion loss on the bi-phase reflection type attenuator. Referring to **Figure 2.2.2.3.3**, the insertion loss state of the attenuator is close to 0.7 dB at a minimum impedance of 2 Ω , and it is close to 0.5

dB at a maximum impedance of 1.6 k Ω . The minimum insertion loss state of the attenuator in each branch of the vector modulator sets a limit to the minimum insertion loss circle of the vector modulator. The minimal insertion loss circle of the vector modulator is taken to be the worst case 0.7 dB for guaranteed symmetric operation. This means that if the modulating vector was to be kept at a constant maximum amplitude and rotated 360° to form a circle in the I/Q-plane, the maximum amplitude of the vector is 0.7 dB down from the 6 dB minimum theoretical insertion loss of the vector modulator.

Figure 2.4.1.1.1 – RF Impedance Characteristics of PIN Diodes as a Function of Forward Bias Current

Other factors in practice that will affect the minimum insertion loss of the vector modulator are, insertion loss of couplers, insertion loss of transmission lines and matching conditions. Typical insertion loss the couplers used in the prototype are around 0.23 dB. For the purpose of the following analysis, the minimum insertion loss of the reflection type attenuator is rounded to 1 dB, which closely resembles the measured performance of the vector modulator implemented for the FFCS prototype.

The isolation of the 3 dB quadrature hybrids used in the reflection type attenuators sets the maximum insertion loss state achievable by the attenuators. In an ideal coupler, the isolation is infinite, which means there is no signal leaking from the input port to the isolated port of the attenuator (feed-through from input to output). Typical isolation

specifications for 3 dB hybrids are around 20 dB. The isolation of the quadrature hybrid used in the FFCS prototype is around 23 dB.

Figure 2.4.1.1.2 – Restricted Vector Modulator I/Q-Plane Coverage

The dynamic range of the attenuator can be interpreted as the isolation of the 3 dB quadrature hybrid coupler less the minimum insertion loss state of attenuator defined by the PIN diodes minimum impedance. In the case of the prototype's reflection type attenuator, the dynamic range would be $(23 \text{ dB} - 1 \text{ dB}) = 22 \text{ dB}$. The limited

attenuation dynamic range restricts the possible modulation states of the vector modulator. The practical coverage of the vector modulator $S_{21_{\text{vm}}}$ under conditions of limited attenuator dynamic range is illustrated in **Figure 2.4.1.1.2** using equation (2.2.2.2.3.3). The maximum and minimum attenuation states of the I and Q components are $|S_{21_{\text{max}}}|$ and $|S_{21_{\text{min}}}|$. The minimum insertion loss circle of the vector modulator is also shown in the figure. The circle has a radius equal to the magnitude of the resultant vector corresponding to the one component fully attenuated up to $|S_{21_{\text{min}}}|$ and the other in its minimum insertion loss state of $|S_{21_{\text{max}}}|$. The minimum insertion loss state of the vector modulator is close to **7 dB**.

Figure 2.4.1.1.2 shows that if the vector modulator is required to modulate near the pure-I or pure-Q axes, these regions are unattainable. The pure-I and pure-Q components cannot assume a transmission coefficient of less than $|S_{21_{\text{min}}}|$. Looking at the vector plot shown in **Figure 2.4.1.1.2**, the maximum insertion loss state of the vector modulator is actually $\sqrt{2} |S_{21_{\text{min}}}|$, and corresponds to the four possible states where both branch components are at their maximum attenuation state $|S_{21_{\text{min}}}|$. The dead-zones around the axes directly impact the achievable cancellation performance of the FFCS in the case where the vector modulator is required to modulate near the I and Q axes.

Unachievable I/Q vector modulation requirements can be interpreted as phase error and amplitude imbalance as a function of the required modulation. To observe the impact of the dead-zones on cancellation performance, the required modulation is assumed to be the variation of the pure-I component with the pure-Q component ideally completely attenuated. Hence the phase error is taken as the phase difference between the resulting vector modulation and the required I-axis component modulation. An expression for the phase error as a function of the required ideal pure I component transmission coefficient $|S_{21_{\text{vm}}}|$ is derived from the vector diagram of **Figure 2.4.1.1.2** and is expressed as equation (2.4.1.1.2) in units of degrees ($^{\circ}$). The amplitude imbalance is interpreted as the difference in magnitude between the practical (physically realizable) vector modulation transmission coefficient and the required (ideal) I-axis component modulation

transmission coefficient. An expression for the amplitude imbalance as a function of the required ideal pure I component transmission coefficient $|S21_{vm}|$ is derived and expressed as equation (2.4.1.1.3) in units of (dB). The derived equations assume the required vector modulation is kept within the dynamic range of the vector modulator. This means the required I-axis component is kept within $|S21_{min}|$ and $|S21_{max}|$.

$$\phi_{err}(S21_{vm}) = \tan^{-1} \left(\frac{|S21_{min}|}{|S21_{vm}|} \right) \quad (2.4.1.1.2)$$

$$\Delta A_{err}(S21_{vm}) = 20 \cdot \log \left[1 + \frac{\sqrt{(|S21_{vm}|)^2 + (|S21_{min}|)^2} - |S21_{vm}|}{|S21_{vm}|} \right] \quad (2.4.1.1.3)$$

Given that the practical vector modulator (limited modulation constellation) is used to modulate a signal, **Figure 2.4.1.1.3** shows a graph of the phase error and amplitude imbalance that would result between the original signal and the modulated signal as a function of the level of attenuation required along the pure I-axis (modulation is in the dead-zones). The attenuation is the required I-axis component modulation less the minimum insertion loss state of the practical vector modulator ($|S21_{vm}| - |S21_{max}|$). The maximum phase error and amplitude imbalance occur when the vector modulator is required to modulate near the origin of the I/Q-plane where both the I and Q components must be attenuated significantly (i.e. near $|S21_{min}|$). If the operating point of the vector modulator is maintained near the minimum insertion loss circle (i.e. near $|S21_{max}|$) shown in **Figure 2.4.1.1.2**, the phase error can be kept at a minimum and the amplitude imbalance can effectively be reduced to a point where it is negligible in the cancellation performance compared to the phase error (refer to **Figure 2.1.3.1**).

Figure 2.4.1.1.3 – Resulting Phase error and Amplitude Imbalance as a Function of the Required Pure I-Component Modulation

Figure 2.4.1.1.4 – Cancellation Performance as a Function of Required Pure I-Component Modulation

The cancellation performance limitations is plotted as a function of the amplitude imbalance and phase error resulting from a required modulation in the dead-zones.

Figure 2.4.1.1.4 was produced using equation (2.1.3.1) where α_{dB} and θ_{err} were substituted by the results of equations (2.4.1.1.2) and (2.4.1.1.3) respectively. **Figure 2.4.1.1.4** shows that the worse case limitation in cancellation performance that may occur with a vector modulator with reflection type attenuators is effectively the same as the I and Q branch attenuator dynamic range used in the vector modulator. The maximum achievable cancellation performance can only be achieved in operating conditions that allow the vector modulator to operate within the reachable area bounded by the minimum and maximum insertion loss states of $|S_{21_{max}}|$ and $|S_{21_{min}}|$ respectively.

In practice, the insertion loss in the main path section from the point of sampling to the point of cancellation will be slightly different than that of the cancellation path. The insertion loss may be higher or lower at different frequency components across a given bandwidth due to amplitude ripple of the different components used. The vector modulator will have to compensate in amplitude to assure balance between the main path and the cancellation path at the center frequency. A certain amount of “play” is required to be left to the vector modulator to ensure full 360° operation. The region of attenuation range available can be visualized as an area in **Figure 2.4.1.1.2** bounded by the minimum insertion loss circle and a circle with a radius equal to the minimum insertion loss state less the required extra attenuation range ($|S_{21_{max}}| - \Delta Att_{range}$). Ideally, the average operating point of the vector modulator is set at an amplitude level equal to half the attenuation range, down from the minimum insertion loss state ($|S_{21_{max}}| - \Delta Att_{range} / 2$). Furthermore, in an application where the guaranteed level of cancellation performance required is lower than the dynamic range of the vector modulator, the remainder of the difference between the dynamic range of the vector modulator and the required level of guaranteed cancellation performance can be used as adjustable gain for the cancellation loop.

For example, assuming an FFCS using the vector modulator analyzed in this section requires a guaranteed cancellation performance better than 20 dB, and the vector modulator is subject to conditions that cause it to operate along the pure I or pure Q axes,

(i.e. $\Delta\alpha_{\text{err}} = 0.043$ dB and $\phi_{\text{err}} = 5.7^\circ$), referring to **Figure 2.4.1.1.4**, 2 dB is left as attenuation dynamic range. The attenuation range could be used to compensate for an amplitude ripple of ± 1 dB at the cancellation center frequency.

2.4.2 – Delay Match of Reference Input to I/Q Demodulator

The I/Q demodulator performs an effective signal correlation to detect the presence of the pilot tone components in the cancellation residual error signal. Correlation is a common detection method used to determine the degree of similarity between the two signals. It is an extremely important concept that is widely used for signal processing applications in radar, sonar, digital communication, electronic warfare, and many others [51].

In the FFCS, the I/Q demodulator only yields relative information between the I and the Q components. The exact magnitude information of the I and Q component cannot be directly used as a control signal for the vector modulator and is therefore not necessary. Only the fact that a component is present, yields enough information to reveal the presence of that component left in the cancellation residual error signal.

Resolving the I and Q components of the residual error signal requires a reference plane. The reference plane is set by the LO drive signal to the I/Q demodulator. The transfer function from the vector modulator to the I/Q demodulator was established in **Section 2.2.4**. The reference signal must be aligned in time and phase. In the ideal case of a demodulation process done with reference to a pure sinusoidal tone, time alignment of the two signal being correlated is not important, the process may even be considered wideband. However, in RF design practice, the phase constant of the transmission medium is a critical issues that must be considered.

In the case where the reference signal may be modulated, time alignment is critical. The reference signal may be modulated if the reference is taken to be the noise itself or in a condition where the pilot is stepped to change channel. Furthermore, if a low power

spread spectrum pilot tone was used, time alignment would be required to be less than a fraction of the period of the center frequency to achieve good processing gain.

For the same reasons demonstrated in **Section 2.1.4**, two signals assumed identical (amplitude, phase, frequency, etc...) propagating through two different paths (with the same phase constant), and arriving at a common point in space, will have differing relative phases between the different frequency component if the electrical delay between the two paths is not the same. In the case of two pure sinusoidal tones, aligned in phase with each other at a given frequency, if the electrical length of the two paths employed by the two signals is different, changing the frequency of the tone causes a phase difference between the two signals. The degree with which the phase varies with a change in frequency is a function of difference in electrical length. In the context of the FFCS, this implies that the I and the Q components of the vector modulator and I/Q demodulator become coupled.

Figure 2.4.2.1 illustrates the effect of delay mismatch on error signal decomposition. The figure assumes the two signal paths are the cancellation path and the pilot tone reference path. The vector modulator is coupled directly to the I/Q demodulator. The case illustrated is where the electrical delay, through the cancellation path is longer than that of the pilot tone reference path to the I/Q demodulator RF and LO input ports. The vector modulator input is shown in **Figure 2.4.2.1 (a)** along with the corresponding I/Q demodulator output shown in **Figure 2.4.2.1 (b)** at a given frequency f_c . At frequency f_c , the I/Q quadrants of the vector modulator is aligned with that of the I/Q demodulator. In a transmission line of a fixed length, the phase constant is a function of frequency. As the frequency is increased, considering the relative phase shift of the vector present at the end of the transmission line, a clockwise rotation is observed. It can be said that the number of waves inside the transmission line are packed closer together due to shortening of the wavelength. Vice-versa for a frequency decrease. Therefore, if the two paths have a different length, the change in the number of wavelength that are packed closer together will be different for a given change in frequency. This implies the I/Q demodulator is not

aligned with the vector modulator at all frequencies. The I/Q demodulator output for a positive frequency deviation is illustrated in **Figure 2.4.2.1 (c)** and negative frequency deviation in **Figure 2.4.2.1 (d)**. Although the vector modulator indicates a pure I component, at higher frequencies than f_c , the effective I/Q plane at the vector demodulator is rotated by an angle $\Delta\theta$. The I/Q demodulator now detects a different vector than expected, according to its reference plane.

Figure 2.4.2.1 – Effect of Delay Mismatch on I/Q-Plane Alignment

Delay mismatch in the pilot tone reference path does not necessarily significantly affect the outcome of the convergence. The main effect of small delay mismatch is an increase in convergence time, due to the coupling between the I and the Q components resulting from relative I/Q plane reference rotation. The control system will continue to adapt itself until the I and Q components of the cancellation signal are adjusted such that the error signal is minimized. As long as the angle $\Delta\theta$ remains smaller than $\pm 90^\circ$ (delay mismatch is within $\frac{1}{4} \lambda$), the control system will eventually converge. Although a

significant amount of energy may be “erroneously” transferred to the quadrature component, the system converges because there is still a gradient component pushing to reduce the energy in the loop. As long as the angle $\Delta\theta$ remains smaller than $\frac{1}{4}\lambda$, the negative gradient estimate ϵ will have a dominant component pointed in the direction of maximum cancellation level (considering the MSE surface). In this case, the loop remains stable and converges as decaying exponential with time. Energy is continuously transferred between the I and Q components with some decay. The negative gradient estimate as observed at the output of the I/Q demodulator essentially spirals toward 0 in the sense of $-\Delta\theta$ (i.e. if τ_{ref} is too short a delay, the error gradient rotates in the clockwise direction). The decay would be observed as a downward spiraling curve on the surface plot of **Figure 2.2.3.1**. If the delay mismatch is such that the I/Q plane relative rotation is $\pm 90^\circ$, oscillation results and the loop becomes unstable if $\Delta\theta$ exceeds $\pm 90^\circ$ due to positive feedback.

If the pilot tone was to be modulated, phase dispersion would reduce the correlation bandwidth in proportion with increasing delay mismatch. The reduced correlation bandwidth would increase cancellation performance convergence time since less energy is available to effectively “push” on the integrator towards convergence.

The delay match of the signals arriving at the I/Q demodulator essentially sets the tuning range of the FFCS. If the pilot tone frequency is driven outside the tuning range, the FFCS falls out of lock and becomes unstable. The condition for stable operation is expressed in equation (2.4.2.1). λ_{err} is the difference in the number of wavelengths (at frequency f_c) between the two paths. The factor f / f_c is the normalized frequency deviation from the center frequency f_c at which it is assumed the I/Q plane of the vector modulator is aligned with the I/Q plane of the vector demodulator ($\Delta\theta = 0^\circ$). The tuning range is defined as the range of pilot tone frequencies f where the condition of (2.4.2.1) is satisfied.

$$2\pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c}\right) < \frac{\pi}{2} \quad (2.4.2.1)$$

2.4.3 – Signal to Noise Ratios at Different Nodes in the FFCS Architecture

The maximum cancellation level that can be achieved is limited by the noise floor of the cancellation path. It is important to establish the signal and noise levels at all the nodes in the FFCS to understand the limitations of the system.

Figure 2.4.4.1 illustrates the signal and noise levels at different nodes of the FFCS.

Each node is associated with a typical spectrum representation of the signals present at that node. The lowest level of the spectrum is the thermal noise floor (TNF). The pilot tone is represented by a single frequency component. The (Noise) signal is represented by a uniform spectrum which sits just below the maximum level of the pilot tone (3 dB lower). The noise floor (NFL) at different nodes is indicated by a uniform spectrum which sits above the thermal noise floor.

A few assumptions were made for this simple signal and noise analysis. The coupling values of the directional couplers are indicated beside each coupler. The through path insertion loss of the four directional couplers in the main path are ignored. The noise figure of the PIN diodes which is expected to be related to the noise figure of the baseband section is also ignored (however is significant and should be taken into account in practice). The main path and reference path delay lines are assumed to have an insertion loss of 3 dB. The RF gain blocks in the cancellation path are assumed to have a gain of 16 dB with a noise figure of 3.5 dB. The RF PI attenuator in the cancellation path has an attenuation of 6 dB and the vector modulator is assumed to have an attenuation of 9 dB.

The important point to realize is that the active components in the cancellation path add noise to the cancellation signal. The noise floor of the cancellation path is higher than the thermal noise floor and therefore carries an overall noise figure. In the case of the undesired signal band, the signal to noise ratio in the main path, after the point of cancellation, is reduced in part by the noise figure of the cancellation path and by the cancellation process. The noise figure of the cancellation path sets the dynamic range of the FFCS. The dynamic range can be expressed as the signal to noise ratio of the cancellation signal. The maximum level of cancellation that can be achieved is therefore limited by the noise floor of the cancellation path. The two main contributors to the noise figure of the cancellation path is the noise figure and linearity of the RF gain blocks and the noise in the baseband section that modulates the PIN diodes of the vector modulator.

Appendix B3 demonstrates sample dynamic range calculations for the RF gain blocks. If the signal level at the input of the gain blocks is too high, the IMD products decrease the SNR at the output of the amplifier. The SNR at the output of the amplifier directly limit the maximum level of cancellation performance that can be achieved. **Figure B3.3** shows the SNR at the output of the amplifier as a function of the input signal power level. A typical signal in the vicinity of -29 dBm at the input of the second RF gain block **G2** will have an output signal to noise ratio of approximately 73 dB. The maximum cancellation level that can be achieved is therefore limited to 73 dB because of the linearity of the RF gain blocks.

Assuming the signal levels are kept within the spurious free dynamic range of operation of the RF gain blocks, the cancellation performance is then limited by the noise floor of the cancellation path. In **Figure 2.4.4.1**, the detection dynamic range is calculated at 136 dB. Ideally, the I/Q demodulator could detect and decompose the error signal in the 136 dB dynamic range above the noise floor. However, the I/Q demodulator has a limited sensitivity. The actual sensitivity of the I/Q demodulator is difficult to determine due to a near-DC component being demodulated. The demodulated signal is lost below the flicker noise of the baseband outputs.

The noise in the baseband section can be reduced by narrowing the loop bandwidth. The loop bandwidth is reduced by increasing the integration time of the loop filter or decreasing the loop gain. However, decreasing the loop gain increases the steady state error and hence limits the cancellation performance. Dynamic performance and Cancellation performance are therefore tradeoffs of each other.

Chapter 3 – Hardware Prototype Implementation and Validation of Individual System Function Blocks

3.0 - Introduction

Each function block of the FFCS was designed with specific performance requirements. The performance requirements of each function block were determined following the initial target specification outlined in **Section 1.5** and the theoretical analyses of **Chapter 2**.

Based on the target performance requirement of each function block, suitable hardware components were chosen for the hardware prototype. The schematic capture and PCB layout of the prototype was done using Agilent's Advanced Design System. Final hardware layout vs. schematic (LVS) simulations of the system were not performed in ADS because of limited availability and complexity of hardware models supplied by the manufacturers for the chosen components. ADS was used for the hardware design because of its good "layout vs. schematic" synchronization capabilities. This feature assured all networks were connected properly and eased debugging potential bad connections in the layout.

The hardware prototype was designed as a proof-of-concept rather than an optimized, product oriented design. A preliminary prototype board was populated to evaluate the functionality of the individual system blocks. The preliminary prototype board was equipped with strategically placed access test points between all subcircuits to isolate and probe the individual subcircuits to validate the I/O characteristics. The proof-of-concept FFCS prototype was designed to operate in the frequency band spanning 824 MHz to 849 MHz with a center frequency of 836.5 MHz.

3.1 – FFCS Hardware Component Selection

Figure 3.1.1 – FFCS Schematic

The FFCS hardware schematic is shown in **Figure 3.1.1**. The system blocks are explained in the following sections. The prototype components were chosen to be discrete components mounted on a single sided printed circuit board. A discrete implementation of the prototype was chosen over an RFIC design because of the rapid turn around time of a PCB implementation, ease of testing and low cost of implementation at the initial stage of this proof of concept. The wide selection and availability of pre-built surface mount components operating in the frequency range of the FFCS could essentially be inserted in the design in a “drop-in” fashion to accomplish the system level functions.

3.1.1 – Main Path

Figure 3.1.1.1 shows the schematic of the main path along with the three RF signal ports to the FFCS. The pilot tone signal coupler **CP1** that couples the pilot tone into the main

The 20 dB coupler exhibit a coupling factor of $21 \text{ dB} \pm 2 \text{ dB}$ over the operating range. The 10 dB couplers exhibit a coupling factor of 10 dB with a variation $\pm 1.5 \text{ dB}$ over the operating range. The variation in coupling factors across the frequency band of operation places a limit in the cancellation bandwidth of the FFCS due to amplitude ripple in the couplers. Assuming perfect delay matching and perfect phase match across the entire operating frequency range, a coupling amplitude ripple of $\pm 1.5 \text{ dB}$ limits the cancellation performance to no more than 15 dB near the band edges.

3.5mm side-mounted 50Ω SMA Connectors (82 SMA-50-0-1 / 111 NH) from Huber and Suhner are used as interface ports to the PCB. The 3.5mm 50Ω interface connectors are

standard for wireless laboratory measuring equipment and allow reliable and repeatable interconnections between equipment RF I/O ports.

The input return loss (S11) of the RF input port and pilot tone input port, and output return loss (S22) of the RF output port, were measured using a network analyzer and found to be better than 10 dB over the frequency range from 800 MHz to 900 MHz. 10 dB of return loss causes 10% of the power incident to the port to be reflected. The return loss measured is considered acceptable for this system at the stage of proof of concept since the input-output characteristics of the system do not affect the system performance besides the insertion loss. More care should be taken to match the couplers and transmission lines to achieve better return loss specifications.

The insertion loss of the main path between the RF input port (1) and the RF output port (2) was measured using a network analyzer and found to be 2.7 dB, which is close to the calculated insertion loss of 3 dB from (2.2.1.1). The main path insertion loss results from the loss in the microstrip transmission lines connecting the couplers, loss in the main path coax delay line and the insertion loss of the directional couplers in the main path. 3 dB insertion loss in the main path would reduce the output power of the PA by 50%, which is not desirable.

Referring to **Appendix B2**, the main contributor to the insertion loss are the 10 dB directional couplers, each coupling 10% of the incident power out of the main path. Reducing the coupling value of the directional coupler would reduce the insertion loss in the main path. The other two contributors of insertion loss in the main path are the delay line length and the length of microstrip transmission lines used from the RF input to the RF output between the couplers and interface ports of the FFCS. The insertion loss would be reduced by minimizing the circuit size, hence minimizing microstrip transmission line interconnects between the couplers and interface ports. Provided the design of the vector modulator could also be reduced, the delay line could also be shortened, lowering its contribution to the main path insertion loss.

The insertion loss in the RF paths could also be lowered by using a dielectric material with lower loss tangent and conductors with higher conductivity. Taking into account the effect of delay mismatch on cancellation performance, an application requiring lower cancellation performance may permit to leverage delay mismatch between the main path and the cancellation path to reduce the length of the main path delay line. For example, assuming the delay in the cancellation path is brought down to one λ , referring to **Figure 2.1.4.3**, an application that requires a cancellation performance of 40 dB in 5 MHz (at a center frequency of 836.5 MHz) would not require a main path delay line to achieve the performance goal (assuming no amplitude imbalance and no phase error). In the case where the design of the FFCS would be miniaturized, the insertion loss of the FFCS could theoretically be brought down to the through-path insertion loss of the four directional couplers in the main path, (assuming high coupling values are used, negligible power is coupled out of the main path), which would be approximately 1 dB considering the directional couplers used in the FFCS. For example, the coupler **CP1** could be reduced to 30 dB. CP1 would then couple a negligible 0.1% of power out of the main path, assuming the pilot tone power can be increased by 10 dB. The coupling value **CP3** could be increased to 30 dB assuming the 20 dB increase in insertion loss of the cancellation path can be compensated without significant cancellation path SNR degradation by introducing more gain. The coupling value of **CP4** can be lowered at the expense of cancellation performance due to limited sensitivity of the I/Q demodulator. The coupling value of **CP2** can be lowered at the expense of degrading the SNR of the cancellation path (increasing the noise floor), assuming the insertion loss can be compensated by equivalent gain with acceptable cancellation path SNR degradation.

Assuming the FFCS design was optimized as mentioned above, with no main path delay line and using 30 dB directional couplers in the main path, the insertion loss would be reduced to ~ 1 dB. The cancellation performance would be limited to approximately 40 dB in a bandwidth 5 MHz (assuming other disturbances such as amplitude imbalance and phase error ripple between the signal paths).

The FFCS is designed for a 0 dBm pilot tone signal injected in the RF pilot tone input port (3). The pilot tone signal is split with an Anaren 4D1304 in-phase 2-way Wilkinson power divider **CP9**. The in-phase power divider **CP9** operates in the 800 MHz to 1 GHz frequency range. The pilot tone signal at the outputs of the power divider is reduced by 3.1 dB. One output of **CP9** is fed into the coupling port of the pilot tone injection coupler **CP1**. The other output of **CP9** goes to drive the LO port of the I/Q demodulator.

3.1.2 – Cancellation Path RF Gain Blocks

Figure 3.1.2.1 Cancellation Path RF Gain Blocks

A few important characteristics must be taken into consideration when choosing suitable gain blocks for the cancellation path. Low distortion in the band of interest is important to keep a reasonable correlation between the main signal and cancellation signal. If the cancellation signal becomes too distorted, it will no longer be a replica of the main signal. The phase or amplitude adjustments will not be sufficient to correct the cancellation signal. The level of spectral re-growth in the amplified signal effectively sets the noise floor of the cancellation path in the spectral re-growth bandwidth. **Section 2.1** demonstrated the importance of the maintaining the amplitude balance and phase error between the frequency component of the main signal and cancellation signal. The same rules apply for every component across the entire noise bandwidth to be cancelled. The FFCS consequently requires an amplifier with low amplitude ripple and linear phase vs. frequency characteristic.

The RF gain blocks **G1** and **G2** chosen to amplify the sampled noise in the cancellation path are shown in **Figure 3.1.2.1 (a)** and **Figure 3.1.2.1 (b)** respectively. MSA-2643 silicon RFIC amplifiers from Agilent Technologies are used. The two gain blocks use the same design, which is based on the suggested design procedure contained in the datasheet for the MSA-2643 (see **Appendix C1** for design notes). The RF input and output ports are closely matched to 50Ω , with return loss better than -16 dB and require no extra matching circuitry. The two coupling capacitors **C1** and **C2** are DC blocking capacitors that provide a low RF impedance of about 4Ω at 836.5 MHz. The bias networks formed by the RF choke **L1**, bias resistor **R5** and high frequency decoupling capacitor **C3** sets the bias current of the amplifiers to 42 mA. The resistor **R5** is required to assure thermal stability of MSA series of amplifiers. Capacitor **C3** is a bypass capacitor used to couple any noise present on the supply node to ground. The inductor **L1** is an RF choke that exhibits high RF impedance around 836.5 MHz.

The gain blocks each have a power gain of 16.2 dB with a noise figure of 3.8 dB. The combination of the two amplifiers in the cancellation path provide more gain than is required to overcome the coupling factors of the noise sampling coupler, cancellation coupler and the insertion loss of I/Q vector modulator. The P1dB of the gain blocks is specified at +11 dBm and the OIP3 is specified at +24 dBm. For a main path noise level of -23 dBm to be cancelled, the signal power level input to the first gain stage is around -33 dBm. The dynamic range of the gain block at an input level of -33 dBm is approximately 80 dB (refer to **Appendix B3** for derivation). The signal to noise (or rather signal to distortion) ratio at the output of the amplifier is approximately 80 dB. The signal at the output of the vector modulator may be in the vicinity of -23 dBm leaving an output SNR of 60 dB. The dynamic range of the amplifier at a given input power level is limited by the noise floor or the linearity of the amplifier. If signal power input to the amplifier is too high, the 3rd order intermodulation (IM3) products rise to a level higher than the noise floor. The SNR at the output of the amplifier is consequently degraded. The signal to noise ratio at the output of the cancellation path ultimately sets a

limit on the peak cancellation performance. Hence, in the case of the FFCS, the maximum cancellation level that can be achieved is expected to be in the range of 60 dB.

One gain block was inserted before the vector modulator and the other after the vector modulator. The configuration was chosen to avoid driving the PIN diode reflection type attenuators at a higher level than necessary such that any signal distortion caused by the PIN diodes is kept to a minimum. Further more, the insertion loss of the vector modulator reduces the signal level before being amplified to reduce the distortion effects of the second gain block. If the second gain block was cascaded immediately after the first gain block, the output signal to noise ratio of the two stages would be in the proximity of 50 dB. The cancellation performance would then be limited to 50 dB.

3.1.3 – Cancellation Path Resistive π -Attenuator

Figure 3.1.3 – Cancellation Path Resistive π -Attenuator

The gain blocks used in the cancellation path provide a few more dB of gain than are actually required. The resistive π -attenuator shown in **Figure 3.1.3** is used to adjust the gain in the cancellation path such that the dynamic range of the vector modulator is optimized. The design of the π -attenuator is done as the last fine-tuning step once a good idea of the insertion loss through the main path including the delay line is known. The π -attenuator provides attenuation and broadband match simultaneously. Since it is a passive component, it has negligible distortion effects. Once the π -attenuator value is set, no further adjustment is required.

It was shown in **Section 2.4.1** that to minimize the amplitude imbalance and phase error when the vector modulator is required to operate near the I or Q axes, the I and Q branch of the vector modulator must have the lowest attenuation possible (i.e. operate far from the I/Q-plane origin). Optimizing the dynamic range of the vector modulator is only possible if the amplitude imbalance between the main signal and cancellation signal is already minimal (ideally within a few fractions of dB). Since the gain blocks already provide a few dB more gain than required as a design margin, it is easy thereafter to bring the gain of cancellation path at the optimum value using the π -attenuator.

When the gain is optimized, it is expected that the insertion loss of FFCS through the cancellation path only, with the vector modulator operating anywhere in the polar plane but in its minimum attenuation state, will be the same as the insertion loss of the FFCS through the main path only.

A few steps are required to design the π -attenuator. First, the insertion loss (S21) of only the main path is measured by setting the vector modulator in the cancellation path to its maximum attenuation setting. The cancellation path is then reconnected, the main path delay line is then removed and proper terminations are installed on the main path delay line ports such that no signal passes through the main path. The operating point insertion loss of the cancellation path may then be measured. The cancellation path characteristics can then be obtained with no signal interaction in the cancellation coupler. The S21 measurement of the cancellation path is taken with the π -attenuator bypassed with a straight-through short. The insertion loss in the cancellation path is then lowered by adjusting the π -attenuator such that the operating point insertion loss of the cancellation path is the same as that of the main path.

The practical minimum insertion loss state of the vector modulator must first be determined to set operating point insertion loss of the cancellation path. The vector modulator must be set near its minimum attenuation state $|S_{21}_{\max}|$. In practice, the reflection type attenuators have different insertion loss at their two minimum attenuation

states (refer to **Section 2.4.1**). The worst case insertion loss of the 4 possible states of the I and Q branch operating at 0° and 180° must be found and used as the lowest attenuation state $|S_{21_{\max}}|$ that the vector modulator is capable of producing. The vector modulator is therefore not expected to operate at modulation levels above its lowest attenuation state $|S_{21_{\max}}|$ to guarantee full 360° operation with minimal amplitude imbalance and phase error when the vector modulator is required to operate along the I or Q axes (in the dead-zones). The 4 states are found by alternately setting the control voltage of the reflection type attenuators at 0V and 3V (maximum output of PIN diode driver buffer) while the other branch not being tested is set at its maximum insertion loss state $|S_{21_{\min}}|$. The voltage level for maximum insertion loss state of the reflection type attenuators were found by measuring the transfer characteristics of the individual attenuators, at approximately 625 mV (14 mA current through each PIN diode).

Once the minimum insertion loss state of the vector modulator $|S_{21_{\max}}|$ is found, the amplitude imbalance (difference in insertion loss) between the main path and cancellation path was measure at 7 dB. The result implies that the cancellation path has 7 dB of gain more than is needed to provide a close amplitude balance between the main signal and cancellation signal at the point of cancellation. Keeping a safety margin of about 1 dB to compensate for potential amplitude ripple in the main path and cancellation path, and to be able to observe some performance characteristics of the adaptive control system, such as overshoots and static errors, the π -attenuator is designed to have an attenuation of 6 dB (refer to **Section 2.2.2.2**).

3.1.4 – Cancellation Path Vector Modulator

Figure 3.1.4.1 shows the I/Q vector modulator design. The S''_{noise} signal enters the I/Q vector modulator and is split into two in-phase components that form the I and Q branch. The input signal to the vector modulator is split using an Anaren 4D1304 in-phase 2-way Wilkinson power divider **CP5**. The power divider operates in the 800 MHz to 1 GHz frequency range. The signals at the outputs of the power divider are reduced by 3.1 dB.

One output of power divider **CP5** is fed into the I-branch reflection type PIN attenuator, shown as **Figure 3.1.4.1 b)**, that adjusts the weight of in-phase component of the input signal. The other output of the power divider **CP5** is fed to the Q-branch reflection type PIN attenuator, shown as **Figure 3.1.4.1 a)**, that adjusts the quadrature-phase component of the input signal.

Figure 3.1.4.1 – I/Q Vector Modulator

The signals from the I-branch and the Q-branch are recombined in quadrature using an Anaren 1D1304-3, 3dB quadrature hybrid coupler **CP8**. Q-branch component input to the quadrature hybrid is the quadrature phase component and undergoes a phase shift of 90° at the output port of the coupler. The I-branch component applied to the quadrature hybrid is combined in-phase undergoing no phase shift at the output port of the coupler.

The amplitude of the I-branch and Q-branch are modified independently by their respective reflection type PIN attenuator. The reflection type PIN attenuators use the same type of 3 dB quadrature hybrid as used for signal recombination. The amplitude of the I-branch signal is controlled by the PIN attenuator formed with the hybrid coupler

CP6 while the Q-branch signal amplitude is controlled by the hybrid coupler **CP7**.

The variable impedance is provided with shunt pin diodes on the 0° and 90° ports of the hybrid. Both reflection type PIN attenuators use the same design and each impedance controlled port uses the same bias network with matched PIN diodes to assure as close an RF impedance match as possible between the two terminated ports of the hybrid coupler. The matched PIN diode pairs are Agilent HSMP-386L series general purpose PIN diodes. The HSMP-386L come in a 6-pin package and contain three PIN diodes. Only two of the three diodes in the package are used and one chip is used per attenuator.

The bias network for the PIN diode uses resistor **R9** to limit the current into the PIN diode as the control voltage is varied. The current in each PIN diode was measured to vary between a few μA to $\sim 50\text{ mA}$ as the control voltage is varied from 0V to 3V. The current drive to the PIN diodes is limited by the maximum current drive capability or the maximum voltage swing of the buffer used to drive the reflection type attenuators. RF choke **L2** feeds the bias current to the PIN diode and blocks high frequency RF components from being introduced on the supply line or causing change in the bias level by causing a voltage drop across **R9**. Capacitors **C6** and **C7** are bypass capacitors used to decouple noise from the supply line to ground. The RF path from the hybrid **CP6** 0° and 90° coupled ports is accomplished through the DC blocking capacitors **C4** and **C5**. The choice of DC block capacitors, coupling capacitors and RF chokes was based on the same impedance calculations noted in **Appendix C1**. The result is an effective RF real impedance that can be varied from $\sim 2\ \Omega$ to about $1.6\text{ k}\Omega$. The minimum insertion loss at these two extreme values was measured at roughly 1.5 dB and 1.8 dB respectively.

The maximum attenuation that can be provided by each reflection type attenuator is limited by the isolation between the input port and the isolated port of the quadrature hybrids used. The measured maximum attenuation was approximately 38 dB. The difference between the maximum attenuation and minimum insertion loss states, yield the dynamic range of the reflection type attenuator. Hence the I and Q component can be varied by a maximum of $\sim 36\text{ dB}$. Following derivations of **Section 2.4.1**, this translates

into potential cancellation performance limited to 34 dB, in the case where the vector modulator would be required to operate near the I or Q axes (i.e. I or Q component fully attenuated).

The I/Q balance of the vector modulator is affected by the phase deviation from quadrature between the two branch signal components. The phase error between the outputs of power divider varies with frequency and is typically within 1° at a frequency of 836.5 MHz. The quadrature error in the quadrature recombination hybrid coupler is within 0.1° at 836.5 MHz. The phase reflection type attenuators do not exhibit perfect 180° phase shift between the two phase states. The phase errors vary from part to part and with frequency. It is better to characterize the vector modulator I/Q quadrature error with all the components in place and measure the amplitude and phase characteristic of the prototype implementation as a function of control voltage.

The constellation of the vector modulator is measured by disconnecting the main path delay line and terminating the ports used to connect the delay line in 50Ω . The control voltage on the I and Q reflection type attenuators can be manually adjusted with individual power supplies to observe the constellation points on the network analyzer. The constellation points are obtained by measuring the magnitude and phase of the insertion loss through the cancellation path. The measured constellation of the I/Q vector modulator is not linear. Furthermore, the phase deviation from 180° between the modulation states having a phase 0° - 180° and 90° - 270° , (i.e. the phase error between the two phase states of the voltage variable attenuators) is approximately 9° across the band. The cancellation performance is only slightly affected by the imbalance. The effect of non-linearity and quadrature in the vector modulator impacts the convergence speed of FFCS but does not significantly change the outcome of the convergence for the same reasons as discussed in **Section 2.4.2**, as long as the required modulation resides within achievable constellation points.

3.1.5 – I/Q Demodulator

Figure 3.1.5.1 – I/Q Demodulator Hardware Configuration

The I/Q demodulator hardware configuration is illustrated in **Figure 3.1.5.1**. The I/Q demodulator detects the presence of the reference pilot tone S_{ref} in the error signal S_{error} . S_{error} is decomposed into an I and a Q baseband component using S_{ref} as the reference vector. The level of the I and Q baseband component v_{ei} and v_{eq} are dependent on both the phase correlation between S_{ref} and S_{error} and the power level of S_{error} (refer to **Section 2.2.3** and **Section 2.2.4**).

The I/Q demodulator used is the AD8347 800 MHz to 2.7 GHz Direct Conversion Quadrature Demodulator from Analog Devices, Inc. This demodulator was chosen for its state-of-the-art performance and ease of use. The AD8347 requires minimal extra circuitry to perform the required demodulating function. An RF signal can be down-converted directly to baseband without the use of intermediate frequency and filters. It operates with a single supply of +5 V.

Integrated RF and baseband amplifiers provide 39.5 dB of gain. Target cancellation performance can be achieved without the use of external RF gain blocks introduced in the error path or increasing the coupling value of the error signal sampling coupler.

The quadrature phase error of the I/Q vector demodulator is specified less than 0.8° around 800 MHz at maximum gain. Quadrature phase error does not significantly affect the cancellation performance but may slow convergence for the same reasons discussed in **Section 2.4.2**. The AD8347 has a noise figure of 11dB. The I/Q demodulator can detect RF signals incident on the RF port, below -75 dBm. The distortion effects of the demodulator are negligible in the FFCS since the power of error signal is inversely proportional to the cancellation performance level.

The design for the I/Q demodulator was based on the functional requirements and the suggested designs presented on the datasheet of the AD8347. The three supply connection **VPS1**, **VPS2** and **VPS3** are decoupled with a 100 pF and a 0.1 μ F capacitor. All ground connections on the AD8347 use the same common low impedance ground. The ENBL pin enables the AD8347 when a high level (+VCC) is applied. The enable pin is tied to +VCC to keep the chip enabled.

The LO inputs to the I/Q demodulator **LOIP** and **LOIN** must be driven differentially for optimal performance. A M/A-COM ETC1-1-13 balun is used to convert to the single ended signal S_{ref} to a differential signal. The differential signal is AC-coupled to the **LOIP** and **LOIN** pins of the I/Q demodulator IC via 100 pF coupling capacitors **C15** and

C16. The $200\ \Omega$ shunt resistor across the **LOIP** and **LOIN** pins is used to improve the broadband match to a $50\ \Omega$ source. The RF input of the I/Q demodulator is connected to the **RFIP** and **RFIN** pins. The RF input signal is AC coupled into to the **RFIP** pin and the **RFIN** pin is AC coupled to ground, via $100\ \text{pF}$ coupling capacitors **CP20** and **CP21** respectively. The $200\ \Omega$ shunt resistor across the signal side of the coupling capacitors on the RF input port is used to improve the broadband match to a $50\ \Omega$ source.

The I and Q-channel baseband outputs v_{eI} and v_{eQ} of the I/Q demodulator are differential baseband signals present on the low impedance **IMXO** and **QMXO** pins of the AD8347, whose bias level is the voltage V_{ref} present on the **VREF** pin. V_{ref} is typically $1\ \text{V}$. The AD8347 is equipped with an offset nulling loop that continuously nulls any DC level present on the **IMXO** and **QMXO** pins of the I/Q demodulator. The I and Q-channel offset nulling inputs **IOFS** and **QOFS** pins are used to null the DC offset on the **IMXO** and **QMXO** baseband output pins. However, S_{error} and S_{ref} are at the same frequency, therefore the DC level resulting from the demodulation carries the phase difference information. Correct functionality of the FFCS depends on extracting phase information between the LO and RF signals. The DC level present on the baseband output I/Q demodulator carries important phase information and must not be nulled. Forcing a voltage on the **IMXO** and **QMXO** pins disables the DC offset compensation circuit of the AD8347 and sets the DC bias level on the **IMXO** and **QMXO** pins respectively.

Baseband DC voltage levels corresponding to the RF-LO phase difference depends on temperature, gain and varies from part to part. It is therefore hard to have repeatable performance when tuning this part.

When no RF input signal is applied to the I/Q demodulator, a DC offset bias level may be present on the **IMXO** and **QMXO** pins. DC offsets cause static errors in the convergence of the adaptive control system. The DC offset on the **IMXO** and **QMXO** pins must be reduced to improve the cancellation performance. A resistive voltage divider is used on the **IOFS** and **QOFS** pins to set the bias on the **IMXO** and **QMXO**

baseband outputs of the demodulator. The bias on each output can therefore be manually controlled to null any DC offsets that are present when no signal is applied at the RF input. The $700\ \Omega$ resistors **Rd2** and **Rd4** shown in **Figure 3.1.5.1** consist of a $10\ \text{k}\Omega$ potentiometers in series with $10\ \Omega$ potentiometers **Rd1** and **Rd3** respectively. Coarse and fine control of the bias levels can be realized by tuning the potentiometers to reduce DC offset present on the **IMXO** and **QMXO** pins. The potentiometer allow the offset voltages to be controlled to within a few fractions of a millivolt.

3.1.6 – Baseband Processor

Figure 3.1.6.1 – Baseband Section Schematic

Figure 3.1.6.1 (a) and **Figure 3.1.6.1 (b)** show the schematics of the baseband I and Q channels respectively. The I and Q channel designs are identical. The baseband section closes the feedback loop of the adaptive control system. The controls system consists of a gain stage, an integrator and an active filter.

The baseband outputs v_{eI} and v_{eQ} of the I/Q demodulator are amplified by a differential amplifier. The differential output of the differential baseband amplifier is converted to a single ended buffered signal. The baseband amplification stage is buffered to drive a charge-pump circuit that charges and discharges an integration filter circuit that integrates the baseband error signal. The loop filter at the output of the charge pump is a first order filter which sets the bandwidth of the control loop. An output gain stage, buffers the output of the charge pump. The output buffer has the necessary current drive capability to control the reflection type PIN attenuator via control voltages v_{cI} and v_{cQ} .

3.1.6.1 – Baseband Amplifier

The baseband output of the I/Q demodulator is applied to a differential baseband amplifier. The differential amplifier used is an AD8132 high-speed differential amplifier. Analog Devices Inc. suggests using the AD8132 baseband amplifier instead of the AD8347 integrated baseband amplifier to reduce baseband output offset voltage and noise. The AD8132 is a high-quality differential amplifier compared to the one integrated into the AD8347 I/Q demodulator IC.

The AD8132 baseband amplifier is designed to amplify the baseband output voltage of the I/Q demodulator. The AD8132 makes use of two feedback loops formed with $475\ \Omega$ input resistors and $4.75\ \text{k}\Omega$ feedback resistors. The AD8132 has a closed loop 3 dB bandwidth of about 20 MHz. The input to the I and Q baseband amplifiers is the differential signal between the V_{ref} and v_{eI} , and V_{ref} and v_{eQ} demodulator baseband outputs. The amplified baseband signals become v_{dI} and v_{dQ} .

3.1.6.2 – Differential to Single-Ended Buffer OPAMP

The amplified differential signal v_d is input into the following opamp circuit that consists of a differential input to single ended output buffer. The differential output of the baseband amplifier is converted to a single ended bi-polar voltage with sufficient current drive to operate the charge pump. The amplified single ended output v_{sI} and v_{sQ} are used to drive the charge pump.

The differential to single-ended buffer circuit is built around an AD9631 voltage feedback opamp from Analog Devices Inc. The AD9631 has a slew rate of 1300 V/ μ s which allows rail-to-rail switching speeds above 200 MHz in open loop configuration. The AD9631 amplifier is designed to operate with a closed loop voltage gain of 66 dB. The high gain of the amplifier assures sufficient drive under small signal operation to fully switch the charge-pump. The closed loop 3-dB bandwidth is approximately 100 kHz.

The AC measured performance of the AD9631 indicates a maximum gain of 46 dB and a 1 MHz 3 dB bandwidth. The output swing of the amplifier ranges from ± 3.9 V, which is sufficient to fully switch the charge pump.

3.1.6.3 – Charge Pump and Loop Filter

The charge pump consists of two complementary current sources which either sink current from, or source current to, the charge storage capacitor. The charge pump performs a voltage to current conversion. The charge storage capacitors **C47** and **C45** are used to integrate the charge pump output current. The complementary charge pump is built using two pairs of BJTs. The current source is built with general purpose Philips PNP (BC857BS) double transistor IC. The current sink is built with general purpose NPN (BC847BS) double transistor IC.

The charge pump design is based on the design of an external charge pump using BJTs, presented in [52] (see **Appendix C2** for design notes). The charge pump is designed to sink or source a relatively constant current of 5 mA in and out of the loop filter. The bipolar control voltage v_s applied to the input of the charge pump is in the range of ± 3.8 V. The charge pump has an inverting characteristic. A positive input control voltage activates the NPN current sink and cuts-off the PNP current source. Current is drawn from the loop filter, effectively reducing the charge in the charge storage capacitor. A negative input control voltage activates the PNP current source and deactivates the NPN current sink. Current is pumped into the loop filter, effectively increasing the charge in the charge storage capacitor.

3.1.6.4 – PIN Attenuator Driver

The baseband output stage amplifies the voltage output of the loop filter and provides sufficient current drive to operate the PIN diode attenuator. The input impedance of the PIN diode driver circuit is roughly equal to **R51**. A high input impedance is required to avoid loading the loop filter. Loading the loop filter would cause the loop filter voltage to decay, degrading the steady state error of the control loop. The PIN diode driver circuit has a voltage gain of $A_v = 1.2$ provided by the resistor ratio (**R54** / **R51**) and (**R50** / **R47**). The PIN diode driver is built around the AD9631 voltage feedback opamp from Analog Devices Inc. The 100 Ω resistor used on the positive input terminal is suggested in the datasheet to balance the offset bias current on the differential input terminals of the opamp. Two bypass capacitors are connected between each supply connection and ground as suggested on the datasheet of the AD9631.

3.2 – Printed Circuit Board Layout

The printed circuit board (PCB) layout was performed using the layout tool that is part of Agilent ADS. The PCB layout design was converted to Gerber format and sent to CMR Circuits for manufacture. The prototype PCB is implemented using five copper

conductor layers that have a thickness of 1 ounce copper ($\sim 25 \mu\text{m}$). ParkNelco N4000-13 epoxy laminate and prepreg epoxy system is used for the 4 substrate layers. All vias and PCB traces were finished by emersion silver plating. The N4000-13 resin system is well suited for wireless communication circuits demanding good impedance control, low loss and good thermal stability. The PCB layer configuration is shown in **Figure 3.2.1**.

Copper 1: RF Signal / Baseband Signal	(Appendix D1)
Substrate 1	
Copper 2: RF Ground Plane & Power Ground Plane	(Appendix D2)
Substrate 2	
Copper 3: DC Power + 5 V	(Appendix D3)
Substrate 3	
Copper 4: DC Power - 5 V	(Appendix D4)
Substrate 4	
Copper 5: VM Control Signal / Ground Plane	(Appendix D5)

Figure 3.2.1 – PCB Layer Configuration

Microstrip transmission lines are used to guide the RF signals. The microstrip transmission lines are formed with the RF signal node on the top conductor layer of the PCB with the reference ground plane on the second conductor layer. The RF transmission lines are designed for a characteristic impedance of 50Ω near a signal frequency of 836.5 MHz. 50Ω is the industry standard characteristic impedance used for RF circuit interconnects. The required transmission line width is 25 mil for a dielectric thickness (thickness of the substrate) of 12 mil, dielectric constant of 3.9, copper thickness of $\sim 25 \mu\text{m}$ and loss tangent of 0.014 (refer to **Appendix B1** for derivation). The transmission line width was confirmed and agreed with the **LineCalc** tool that is part of ADS for calculating the required transmission line parameters.

The layout of each conductor layer of the PCB are shown in **Appendices D1 to D5** as indicated in **Figure 3.2.1**. The layout of all subcircuits was done according to suggestions of the individual parts manufacturer. The populated PCB layout is shown in

Figure 3.2.2. Details on the layout of each subcircuit and the parts list are contained in Appendix E.

Figure 3.2.2 – Populated FFCS PCB Layout

The FFCS PCB layout is divided in two sections. The two sections are separated by a wall of vias that are connected to ground. RF interference may be radiated and propagate in the dielectric substrate and interact with the opamp circuits used in the baseband processor, causing DC offsets and noise in the baseband signals. The VIA wall helps to minimize the RF interference introduced from the RF section into the baseband section by providing an effective shield. The wall of vias is made two rows wide with vias separated by $1/30 \lambda_c$, where λ_c is the effective wavelength at the design center frequency. The PCB is accommodated with four tooling holes, one in each corner of the PCB. The tooling holes are used to attach stand-offs to support the PCB avoiding inadvertent contact with conducting objects that could lie around on the work bench. A film of solder mask coats the top surface of the PCB to protect the conducting traces from environmental conditions such as humidity or liquid spills. Varying humidity levels or spilling chemicals on the traces could change the impedance of high-impedance transmission lines or provide a conductive channel between close-by circuit traces. Only landing pads and areas where solder is used to bond the components and test wires have been cleared of solder mask. Solder mask is also used on the under-side of the PCB to protect the baseband signal carrying traces. However, only the signal traces are covered in solder mask under the PCB.

A filled ground plane provides a low impedance path between all the ground nodes in the circuit. A low impedance path between all grounded nodes assures the ground reference of all components in the circuit are at the same voltage potential. The supply connections are also distributed throughout the circuit with separate +5V and -5V filled power planes. Separate filled power planes makes signal routing much easier and low impedance supply feeds.

The bottom conductor layer of the PCB is used both as another ground plane and to route the baseband signals to and from the baseband processor. The differential error signals generated by the I/Q demodulator are routed to the input of the baseband processor and

the output control signal of the baseband processor is routed to the PIN diode bias circuit of the reflection type attenuators.

The PCB Prototype was designed to allow testing of a second FFCS architecture. The second architecture was dropped after preliminary testing could not yield expected results and the FFCS could not enter stable operation. Some layout areas on the PCB prototype are therefore not used and need to be bypassed by short circuits and jumping wires. The areas of the PCB that are not used are marked with an “X box” in **Figure 3.2.2**.

RF Test points were used in the RF section to test the functionality of the individual RF subcircuits. The test bed was implemented on a different PCB. Once all subcircuits were validated, the FFCS was implemented on the final PCB with only the required components to keep the design clean and avoid any parasitic interactions with the unnecessary components. The RF test points were designed using through-hole mounted SMA connector with the signal line connected to the node to be tested through a T-connection as illustrate in **Figure 3.2.3**. The RF test point is removed from the circuit by omitting resistor **R1** and replacing resistors **R2** and **R3** by short circuits. A signal can be applied to, or measured from the node connected to port **P3** by means of the test point **TP1** by removing resistor **R3** and replacing resistors **R1** and **R2** by short circuits. Although resistors are shown in the diagram of **Figure 3.2.3**, a capacitor could also be used instead of one of the resistors to obtain AC coupling from the test point to the node being probed, instead of DC coupling obtained with short circuits. All the resistors, capacitors and inductors used have 0603 form factor (60 mil long by 30 mil wide). Therefore, the resistors and capacitors use the same landing pads dimensions and can be interchanged.

Test points were also used in the baseband section to test the individual subcircuits. The test points in the baseband section are designed by having a through-hole via large enough to solder a small test wire. The via can be isolated from either side of the signal path by removing a short circuit resistor.

Figure 3.2.3 – RF Test Point Configuration (shown disconnected from the RF path)

Figure 3.2.4 – Baseband Section Test Points

3.3 – Delay Line Designs and I/Q Vector Modulator and Demodulator Quadrant Matching

The most technically demanding aspect of the hardware implementation is the precise design of the main path delay line τ_{main} and reference path delay line τ_{ref} . Referring back to **Figure 2.2.1.1**, the main path delay line is adjusted to match the delay through the cancellation path at the center frequency of 836.5 MHz. τ_{main} assures that the main path signal and the cancellation path signal arrives at the same time at the summing junction. τ_{ref} assures that the cancellation residual error signal arrive at the same time as the reference pilot tone signal at the RF port and LO port of the I/Q demodulator respectively. Time alignment of the two signals arriving at the I/Q demodulator is

essential to perform an accurate correlation. τ_{ref} also sets the alignment of the I/Q planes of the vector modulator and the I/Q demodulator.

The importance of precisely matching the delay through the cancellation path was shown in **Section 2.1.4**. Any delay mismatch between the main path and the cancellation path reduces the effective cancellation bandwidth. **Section 2.4.2** discussed the importance of properly aligning the two signals being correlated to obtain a close estimate of the instantaneous error gradient. Improper delay match shifts the effective phase of the error signal with respect to the reference signal, increasing the convergence time of the FFCS control loop and can cause potential instability. The more precision achieved in the design of the delay lines the wider the cancellation bandwidth and tuning range of the FFCS.

Due to the precision required in delay match between the different paths in the FFCS, it is very hard to design the delay lines based on theoretical calculation or hardware simulation. Delay specification is not usually a concern in most standard RF front-end applications. Furthermore, precise delay specifications of most hardware components used could not be obtained by the suppliers. Measuring delays with pico-second precision is very hard to perform in the lab using standard equipment. A network analyzer can simply yield a rough approximation of the actual delay through the RF path measured. Although the measurement may be very accurate, the tolerance variations in the RF cable, and the difference in matching conditions caused by the introduction of the delay line in the RF system may affect the effective delay through the parallel RF path.

The delay lines required in the FFCS were designed manually. The technical procedure devised to perform a precise delay line design is described. Designing the delay line of the reference path τ_{ref} first, eases the precise design of delay line of the main path τ_{main} . The delay lines were designed using Huber & Suhner Sucoform coax cable of type SM 86. SM 86 is hand formable coax cable having a characteristic impedance of 50Ω and a

velocity of propagation of 71% the speed of light in free space. The delay is specified at 4.7 ns/m with an attenuation of about 0.67 dB/m.

An approximate of τ_{ref} and τ_{main} to be designed is obtained by measuring or estimating the delay through sections of the RF paths from the pilot tone input to the two RF ports of the I/Q demodulator and through the main cancellation signal path. The approximation is used to design a delay line to within one wave length above the optimal coax delay length required. From the initial length, fine tuning can be done by hand trimming the coax cable with millimeter precision. To trim the delay line, one end was secured with an SMA connector while the other end was inserted in the connector and held in place by hand to test until the optimized design was achieved, at which point the connector was secured to the coax. The actual hardware measurement of interest to yield an approximate delay is tabulated in **Table 3.3.1**.

RF Path	Delay (ns)
Pilot Tone Input Port to FFCS RF Output Port with VM = (1,0)	5.132
Error Signal Sampling Coupler to RF Port Input of I/Q Demodulator (estimate)	0.445
Total Delay from RF Pilot Tone Input to RF Port of I/Q Demodulator	5.577
Pilot Tone Input Port to Input Port of the Pilot Tone Reference Path Delay Line	0.709
Pilot Tone Reference Path Delay Line Output to I/Q Demodulator LO Port (estimate)	0.022
Total Delay from RF Pilot Tone Input to LO Port of I/Q Demodulator	0.731
Required Pilot Tone Reference Path Delay	4.846
RF Input Port to Main Path Delay Line Input Port	1.010
Main Path Delay Line Output Port to RF Output Port	0.905
RF Input Port to RF Output Port with VM = (1,0)	4.310
Required Main Path Delay	2.395

Table 3.3.1 – RF Path Delay Measurement and Approximation

τ_{ref} is optimized by opening the control loop to allow manual control over the vector modulator. The vector modulator control voltages are set to $V_{cI} = 0$ V and $V_{cQ} = 0.629$ V. The pilot tone reference path delay line was trimmed until the baseband output of the I/Q demodulator was aligned with the vector modulator. As τ_{ref} is trimmed and tested, the I/Q plane of the demodulator (baseband output) rotates counter-clockwise until only

the positive I component is present. Assuming the estimate of the delay line was within one wavelength of the optimal delay, the delay from the pilot tone input port to RF and LO port of the I/Q demodulator should be matched. A quick way to test this condition is to change the frequency of the pilot tone (say from 800 MHz to 1 GHz). The I/Q plane rotation of the I/Q demodulator in the case of close delay match is much less than in the case of imperfect delay match. If the delays are closely matched, the I/Q plane rotation should be well within $\pm 90^\circ$ with a frequency variation of 300 MHz. However, the reader must take note that the phase constant of the two paths is different since the cancellation path contains different couplers and active components. The phase constant of both paths is not the same since the propagation conditions of both paths are different. Therefore, there will always be I/Q plane rotation with varying frequency unless the phase constant of both paths can also be matched. The effect of the phase constant difference is negligible in the range of operation of the FFCS.

Once τ_{ref} is designed, the FFCS is ready for operation. Any delay length can be introduced in the main path and the FFCS will lock, provided that the insertion loss through the main path does not exceed the dynamic range of the cancellation path. One characteristic that can be exploited to fine tune τ_{main} is the cancellation bandwidth as a function of delay match. The control loop of the FFCS is closed by connecting the control loop output to the respective vector modulator control inputs. The main path delay is trimmed until the cancellation bandwidth is brought to the desired performance level.

The final measured delay line values for the FFCS were $\tau_{\text{main}} = 2.70$ ns (554 mm) and $\tau_{\text{ref}} = 4.76$ ns (1007 mm).

3.4 – Completed FFCS Prototype

Figure 3.4.1 – Completed FFCS Prototype

Figure 3.4.1 is a picture of the FFCS being tested in the laboratory. The operating characteristics are summarized in **Table 3.4.1**.

Table 3.4.1 - General FFCS Physical and Operating Characteristics	
PCB Dimensions (incl. connectors, stand-offs and delay lines)	21.36 cm (wide) x 24.33 cm (long) x 10 cm (high)
Supply Connections (two SMA ports referenced to ground)	+5V, -5V and GND (0 V)
DC Power Consumption (Average)	~ 2 W : 300 mA (pos), 100 mA (neg)
RF Inputs (two 50 Ω SMA input ports)	RF in, Pilot Tone Reference in
RF Outputs (one 50 Ω SMA output port)	RF out
Pilot Reference Tone Input Power Level	0 dBm
RF Noise Input Power Level	-25 dBm

Chapter 4 – FFCS Hardware Prototype Performance

4.0 - Introduction

The performance of the hardware prototype was evaluated in the lab through a series of measurements. The FFCS is a 2-port system with one control input port. The system performance was measured and characterized in terms of its scattering parameters (S-Parameters) using an HP8753E Network Analyzer. **Table 4.0.1** shows the general settings of the network analyzer used for calibration and testing.

Table 4.0.1 - General Network Analyzer Settings for Calibration	
Model #	HP 8753E 30 kHz - 6 GHz
Calibration Kit	HP 85052D 3.5mm Economy Calibration Kit
Start Frequency	700 MHz
Stop Frequency	1000 MHz
Number of Points	1601
Test Port Power	-33 dBm
Sweep Time	600 msec
Sweep Type	Linear

Figure 4.0.1 – FFCS Lab Measurement Test Setup

Figure 4.0.1 illustrates the hardware test bench setup around the FFCS to measure the performance. Port 1 (forward) of the network analyzer is connected to the RF input port of the FFCS. Port 2 (reverse) of the network analyzer is connected to the RF output port of the FFCS. The pilot tone is generated using a Marconi Instruments 10kHz – 5.4 GHz

2032 RF Signal Generator. All the RF test signals used as pilot tones can be synthesized using the RF signal generator.

The FFCS requires +5V and -5V supply connections referenced to common ground to operate. The DC Power Supply used to deliver power to the FFCS is an “HPE3620A 0-25 V, 1.0 A, Dual Output DC Power Supply”.

4.1 – FFCS Main Path Transmission Characteristics

The transmission characteristics from the RF input port to the RF output port of the FFCS are measured with the system in passive mode (i.e. system power is off and no pilot tone signal is applied). Passive mode shows the transmission characteristics of the main path when no active components are powered. The FFCS is also measured in inactive mode. Inactive mode is considered as the FFCS powered on but with the vector modulator in full attenuation. Measuring the inactive mode characteristics of the FFCS gives a better demonstration of the main path characteristics.

4.1.1 - Passive Mode FFCS Transmission Characteristics

The S-Parameters of the FFCS were measured for the main path with the FFCS in passive mode. When no power is applied to the FFCS, the RF gain blocks are not powered up and the cancellation path is in high isolation state. The measurements were taken in the frequency range of 700 MHz to 1 GHz to illustrate the broadband characteristic of the design.

Figure 4.1.1.1 (a) and **Figure 4.1.1.1 (c)** illustrate the FFCS input and output return loss parameters **S11** and **S22**. These parameters characterize how well the input and output ports of the system are matched to 50Ω. The **S11** and **S22** coefficients quantify the amount of signal power that is reflected from the input and the output port respectively as a function of frequency. The coefficients are determined by taking the ratio of the

reflected signal over the incident signal to the respective port: $20 \log(v_i/v_r)$ or $10 \log(p_i/p_r)$, where v_i and v_r are the incident and reflected voltage of the signal, and p_i and p_r are the incident and reflected power of the signal. **Figure 4.1.1.1 (b)** and **Figure 4.1.1.1 (d)** illustrate the relative phase of the **S11** and **S22** parameters respectively. The network analyzer is limited in displaying phase angles between $\pm 180^\circ$, hence the 360° jumps that occur when the phase exceeds $\pm 180^\circ$. In **Figure 4.1.1.1 (a)** and **(c)**, the return loss is lower than -11 dB over the design frequency band, allowing less than 8 % of the incident input signal power to be reflected.

Figure 4.1.1.1 – Return Loss Characteristics of FFCS in Passive Mode

Figure 4.1.1.2 – Forward Transmission Characteristics of FFCS in Passive Mode

Figure 4.1.1.2 (a) illustrates the forward transmission parameter S_{21} . The S_{21} represents the insertion loss characteristics of the FFCS in passive mode. The insertion loss is attributed in large part to the power coupled in the 4 directional couplers **CP1**, **CP2**, **CP3** and **CP4** used in the main path and the loss in the delay line τ_{main} . Ideally, the system would be transparent in passive mode such that any signal in a bandwidth of interest passing through it would not be distorted and would incur minimal attenuation. The insertion loss in the main path varies in proportion to the length of delay line used and the sum of the powers that are coupled out of the main path into the coupled ports of the directional couplers. **Figure 4.1.1.2 (b)** shows the linear characteristic of the relative phase as a function of frequency. The phase shift is expected to be linear in a non-

dispersive transmission medium. The delay across the frequency band of interest is close to $4.28 \text{ ns} \pm 20 \text{ ps}$ as shown in **Figure 4.1.1.2 (c)**.

The reverse transmission characteristics **S12** of the FFCS in passive mode is the same as the forward transmission coefficient **S21**. The reverse transmission parameter is the ratio of the signal transmitted from the RF input port (1) over the signal incident to the RF output port (2). The main path is symmetric in passive mode, hence the **S12** parameter is expected to be the same as **S21**. The reverse transmission characteristics of the FFCS in passive mode are shown in **Figure 4.1.1.3**.

Figure 4.1.1.3 – Reverse Transmission Characteristics of FFCS in Passive Mode

4.1.2 – Inactive Mode FFCS Transmission Characteristics

The FFCS is considered in inactive mode when power is applied to the FFCS and the vector modulator is manually set to full attenuation such that the cancellation path is in high isolation mode and no pilot tone is applied to the pilot input port. The cancellation performance measured with reference to the S_{21} insertion loss curve of **Figure 4.1.2.1**, when the FFCS is in inactive mode. The S_{21} characteristic of **Figure 4.1.2.1 (a)** is subtracted from the measured S_{21} characteristic obtained during cancellation to remove the insertion loss characteristic from the cancellation measurement.

Figure 4.1.2.1 – Reverse Transmission Characteristics of FFCS in Inactive Mode

4.1.3 – Pilot Tone Input Port to FFCS RF Output Port Transmission

Characteristics

The pilot tone signal input to the FFCS to control the center frequency of the cancellation notch is injected in the main path. The introduction of a pilot tone in the main path contributes a level of added noise to the system. The return loss characteristic (S11) of the pilot tone input port is shown in **Figure 4.1.3.1**. The return loss measurement shows that the pilot tone input port is not very well matched to 50 Ω . It is suspected that the matching is not very good due to a layout conversion error that caused the through-holes to insert the standing SMA connectors not to be plated through during manufacture of the PCB. Furthermore, the return loss at the LO input of the I/Q demodulator circuit is rated higher than -10 dB. The matching conditions at the pilot tone input port cause between 10% to 20% of the input power to be reflected back to the signal source.

Figure 4.1.3.1 – Pilot Tone Input Port Return Loss Characteristics (S11) (Passive Mode)

The transmission characteristics (S21) of the RF path from the pilot tone input port (3) to the RF output port (2) of the FFCS are shown in **Figure 4.1.3.2**. The transmission

characteristics show that the insertion loss from the pilot tone input port to the RF output port of the FFCS is close to 27 dB and the electrical delay is close to 11 ns.

Figure 4.1.3.2 – Pilot Tone Input Port to RF Output Port RF Path Transmission Characteristics (S_{21}) (Passive Mode)

4.2 – Cancellation Performance Manually Adjusted to Cancel near 836.5 MHz

The FFCS is designed to provide maximum cancellation around a center frequency of 836.5 MHz. The FFCS is first tested by manually adjusting the v_{cI} and v_{cQ} control input voltages on the vector modulator to provide cancellation at 836.5 MHz. Manually adjusting the cancellation provides a measure of the cancellation performance that can be achieved without the limitations from the adaptive control circuit.

4.2.1 – Network Analyzer Measurements of FFCS Manually Adjusted near 836.5 MHz

The v_{cI} and v_{cQ} nodes connecting the PIN diode driver to the PIN control bias circuit of the reflection type attenuator are opened by removing a short circuit that closes the loop of each baseband channel. Probe wires are connected directly from a variable power supply to the control inputs of the vector modulator. The control voltage is then adjusted with precision until cancellation is centered at 836.5 MHz. The performance is observed on the network analyzer. The two variable problem is relatively straight forward to solve by tweaking each control voltage alternately and one at a time as the cancellation performance level is decrease to the limit of the tuning precision of the tuning knob of the power supply.

The cancellation performance characteristics are shown in **Figure 4.2.1.1 (a)**. The cancellation characteristic “V”-shape is centered at 836.5 MHz. The cancellation bandwidth at a cancellation level of 30 dB is 45 MHz, 14.4 MHz at a cancellation level of 40 dB and 4.5 MHz at a cancellation level of 50 dB. Comparing the data with **Section 2.4.1**, the delay mismatch between the main path and the cancellation path appears to be close to $0.2 \lambda_c$, assuming negligible amplitude imbalance and phase error. However, the main path delay line was matched to within $0.02 \lambda_c$ by trimming the delay line to within ± 3 mm from perfect delay mismatch. The bandwidth limitation is not caused by the delay mismatch but rather by the amplitude imbalance that is a function of the frequency. The reverse transmission characteristics S11 of the FFCS remain the same as in **Figure 4.1.1.3**.

Figure 4.2.1.1 – Cancellation Performance Manually Adjusted at 836.5 MHz

The amplitude imbalance between the main path and cancellation path was measured by removing the main path delay line and terminating the main path delay line ports in $50\ \Omega$ and then measuring the characteristics of the cancellation path only. **Figure 4.2.1.2** shows the amplitude imbalance and phase error measured by subtracting the main path characteristics from the cancellation path characteristics. The main path characteristics used are those of **Figure 4.1.2.1**. The difference between the main path parameter and cancellation path parameters was compensated such that cancellation is centered at 836.5 MHz.

Figure 4.2.1.2 – Measured Amplitude Imbalance and Phase Error Between the Main Path and the Cancellation Path

The amplitude imbalance of **Figure 4.2.2 (a)** closely corresponds to the actual coupling characteristics of the couplers **CP2**, **CP8** and **CP4** combined. Any coupling variations as a function of frequency in the couplers cannot be compensated for and directly impact the cancellation performance of the FFCS (see datasheets of the Anaren 1D1304-3 and 1D1304-10 directional couplers for characteristics). The impact of amplitude imbalance on cancellation performance is demonstrated in **Figure 4.2.1.3** where a plot of the measured cancellation performance is plotted along with the calculated cancellation performance as a function of the amplitude imbalance curve of **Figure 4.2.1.2 a)** and both amplitude imbalance and phase error of the curves from **Figure 4.2.1.2 a)** and **Figure 4.2.1.2 b)**. The curves of cancellation performance as a function of amplitude imbalance and phase error are reproduced using (2.1.3.1) assuming negligible delay

mismatch. The delay mismatch can be neglected because the impact of amplitude imbalance on the cancellation performance outweighs the effect of delay mismatch. Furthermore the effect of delay mismatch is included in the phase error measurement of **Figure 4.2.1.2 b).**

Figure 4.2.1.3 – Measured Cancellation Performance and Calculated Cancellation Performance as a Function of the Measured Amplitude Imbalance and Phase Error.

4.2.2 – Spectral Measurements of Manually Adjusted Cancellation Performance Around 836.5 MHz

The FFCS was tested with a noise signal applied to the RF input port. The RF output of the FFCS was connected to a spectrum analyzer. **Table 4.2.2.1** shows the setup of the spectrum analyzer.

Table 4.2.2.1 - Spectrum Analyzer Measurement Settings	
Model #	HP 8563E 9 kHz - 26.5 GHz
Start Frequency	824 MHz
Stop Frequency	849 MHz
Display Type	Maximum Hold
Resolution Bandwidth	30 kHz
Video Bandwidth	1 MHz
Sweep Time	70 ms
Sweep Type	Linear

Figure 4.2.2.1 shows the spectrum of the pilot tone residual signal at the output of the FFCS. The effective noise floor of the spectrum analyzer for the measurement settings used is -89 dBm. The signal power levels measured with the spectrum analyzer with the FFCS in passive mode are used for reference. Only information specific to the relative amplitude difference between signals processed in the FFCS are obtained from the measurements. The pilot tone level is at -28.67 dBm at the output when the FFCS is in passive mode. The pilot tone level reflects the insertion loss from the pilot tone input port (3) to the RF output port (2) of the FFCS as measurement show in **Figure 4.1.3.2 a**).

Figure 4.2.2.1 – Pilot Tone Signal Spectrum at Output of FFCS (Passive Mode)

Figure 4.2.2.2 shows the spectrum of the RF input signal that is used as noise signal to be cancelled by the FFCS. The input RF signal spans a bandwidth of 16 MHz and has maximum components across the bandwidth reaching -45.33 dBm.

Figure 4.2.2.2 – FM Noise Signal Spectrum at the Output of the FFCS (Passive Mode)

Figure 4.2.2.3 shows the output spectrum of the FFCS with cancellation manually adjusted at 836.5 MHz with the 0 dBm pilot tone of Figure 4.2.2.1 applied at the pilot tone input port. The Pilot tone level has been reduced by 42.83 dB.

Figure 4.2.2.3 – FFCS Output Spectrum with 0 dBm Pilot Tone Signal Applied and Cancellation Notch Manually Adjusted for 836.5 MHz.

Figure 4.2.2.4 shows the output spectrum of the FFCS with the cancellation manually adjusted at 836.5 MHz with the 0 dBm pilot tone of Figure 4.2.2.1 and the FM noise signal of Figure 4.2.2.2 applied to the RF inputs of the FFCS. The noise level has been

reduced by a level of 40 dB at an offset of 2.5 MHz from the center frequency, 38 dB at an offset of 5 MHz and 33 dB at an offset of 8 MHz.

Figure 4.2.2.4 – FFCS Output Spectrum with Pilot Tone and FM noise Signal Applied to the RF Input Ports with Cancellation Notch Manually Adjusted for 836.5 MHz

4.3 – Closed Loop Cancellation Performance of FFCS with Control Pilot Tone Signal

The closed loop performance of the FFCS is obtained by connecting the baseband processor output signals v_{cI} and v_{cQ} to the respective I and Q control inputs of the vector modulator. When the power to the FFCS is initially turned on with no pilot tone signal applied, the FFCS is not tuned to cancel a particular frequency. The vector modulator may be sitting at any modulation level or free running. The FFCS must be finely tuned prior to first use.

4.3.1 – Nulling the DC Offsets of the Baseband Processor and Locking the FFCS to 836.5 MHz Pilot Tone

The DC offsets on the **IMXO** and **QMXO** pins vary with temperature. After power is applied to the FFCS, it is good practice to allow a warm-up time of a minute to let the temperature of the ICs settle. A 0 dBm pilot tone applied at 836.5 MHz is used to position the FFCS near the operating point where it will be operated to cancel noise. When the 0 dBm single pilot tone is applied to the pilot tone input control port, the vector modulator is tuned to a particular control state and the FFCS is locked to a frequency in the proximity of the design frequency and may not exhibit the optimal cancellation performance level. DC offsets present in the baseband path, at the baseband outputs of the I/Q demodulator, amplifiers and charge-pump cause the system to converge to an incorrect value. The DC offsets in the baseband path is caused by ambient effects such as temperature and external noise pickup. The first FFCS prototype does not have any DC offset nulling circuitry. The DC offsets are controlled manually by adjusting the **IOFS** and **QOFS** (pins 13 and 16) of the I/Q demodulator. The **IOFS** and **QOFS** pins of the I/Q demodulator control the DC offset present on the baseband I and Q signal outputs **IMXO** and **QMXO**. A variable voltage divider is used to fine tune the DC offset on the **IMXO** and **QMXO** I/Q demodulator baseband outputs by applying a voltage offset on the **IOFS** and **QOFS** pins of the I/Q demodulator (see **Section 3.1.5** for details on the voltage divider configuration).

Once the DC offset on the **IMXO** and **QMXO** are fine-tuned such that cancellation performance is centered at 836.5 MHz, the FFCS is operational. As soon as the pilot tone is turned on, the FFCS locks on the pilot tone and tunes to provide maximum cancellation performance at 836.5 MHz.

Figure 4.3.1.1 shows the cancellation performance of the FFCS with a 0 dBm pilot tone of 836.5 MHz (solid curve) along with the manually adjusted cancellation performance of **Figure 4.2.1.1**. The cancellation performance with a 2.5 MHz offset of the center

frequency is degraded because the sweeping -23 dBm test tone of the network analyzer overlaps with the pilot tone. The test tone of the network analyzer adds with the pilot tone and causes the sampled error signal to be inaccurate near the center frequency. The control loop tries to compensate for the composite error signal that does not accurately represent the actual error in cancellation performance. The FFCS is therefore tuned to an incorrect state that does not result in optimal cancellation performance. Both the FFCS and the network analyzer operate on the basis of a tuned receiver. Furthermore, the non-cancelled portion of the pilot tone adds to the test tone of the network analyzer degrading the measurements near the center frequency. The de-tuning effect caused by frequency components near the pilot tone could be reduced by reducing the control loop bandwidth of the FFCS.

Figure 4.3.1.1 – FFCS Locked to 836.5 MHz, 0 dBm Pilot Tone

4.3.2 – Pilot Tone Control

The center frequency of the FFCS can be tuned at a different frequency by changing the frequency of the applied pilot tone. The adaptive control system causes the FFCS to lock or follow the pilot

Figure 4.3.2.1 illustrate the cancellation performance of the FFCS when tuned at different frequency values. The cancellation performance degrades as the pilot tone frequency is offset from the design band of the FFCS ranging from 824 MHz to 849 MHz. The maximum cancellation performance within the design band appears close to 45 dB. Although the FFCS is capable of operating outside of the design frequency band, when the pilot tone is tuned outside the design bandwidth, the cancellation performance degrades because the amplitude and phase imbalance is harder to control. The operating frequency fall out of the design band of the hardware components and the delay mismatch between the pilot tone reference path and error signal causes significant DC offset at the output of the I/Q demodulator (see **Section 2.4.2**). The offset between the point of maximum cancellation performance and pilot tone frequency is caused by the frequency dependant characteristics of the I/Q demodulator which result in I and Q baseband DC offsets.

Figure 4.3.2.1 – Measured Cancellation Performance at Different Pilot Tone Frequencies

The cancellation performance can be better characterized by using the test signal of the network analyzer as the pilot tone control signal input to the FFCS. When the test signal is used as the pilot tone, the tuned receivers of the FFCS and the network analyzer become decoupled. The same reference signal is used in both systems, hence no parasitic

interaction occur between the FFCS and the network analyzer. The transmission characteristics can be measured from the pilot tone input to the RF output of the FFCS. The cancellation performance can then be measured by subtracting the passive mode transmission characteristics of **Figure 4.1.3.2** from measurement. The cancellation performance at the pilot tone frequency as the pilot tone is swept from 700 MHz to 1 GHz is shown in **Figure 4.3.2.2**. The test signal of the network analyzer was adjusted to 0 dBm, the sweep time was 600 ms.

Figure 4.3.2.2 – Measured Cancellation Performance at Pilot Tone Frequency

Figure 4.3.2.2 shows that the FFCS can only be tuned at pilot tone frequencies above 803 MHz. The lower bound of the tuning capabilities of the FFCS is limited to operating capabilities of the I/Q demodulator which is not rated to operate below 800 MHz and results in the step observed at 803 MHz in **Figure 4.3.2.2**. The cancellation performance shows the design is optimized for maximum cancellation performance at 836.5 MHz. The maximum cancellation level exceeds 50 dB across the design frequency band. The positive cancellation level observed when the FFCS is out of lock below 803 MHz corresponds to the main path insertion loss compensation of **Figure 4.1.2.1 a)** used to extract the cancellation performance with some variation due to the unknown setting of the vector modulator.

4.3.3 – FFCS Dynamic Tuning Performance

The tuning speed performance of the FFCS was also investigated by sweeping the test tone used as a pilot tone at a rate of 3.75 ms in a bandwidth of 25 MHz centered at 836.5 MHz (11 steps of 340 μ s). **Figure 4.3.3.1** demonstrates that the FFCS is able to track the pilot tone signal at this rate.

Figure 4.3.3.1 – Measured Cancellation Performance with Pilot Stepped at a Rate of 3.75 ms

4.3.4 – Spectral Measurements at the Output of the FFCS Trained at 836.5 MHz

The spectrum at output of the FFCS was obtained with the spectrum analyzer using the same settings outlined in **Section 4.2.2**. The spectrum of **Figure 4.3.4.1** shows the FFCS trained at 836.5 MHz by a 0 dBm pilot tone. The cancellation level achieved at the pilot tone frequency is above 38 dB. However, -64.17 dBm spurs are present at 290 kHz offset from the center frequency. The spurs are caused by switching noise in the baseband section. As the error signals is reduced below the sensitivity level of the I/Q demodulator, the amplified noise of the baseband outputs of the I/Q demodulator is high enough to toggle the charge pump state. The noise modulates the I/Q modulator which produces offset spurs from the pilot tone center frequency. The magnitude of the offsets are governed by the loop bandwidth characteristic.

Figure 4.3.4.1 – Spectral Measurement at the Output of the FFCS Trained with a 0 dBm Pilot Tone

Figure 4.3.4.2 shows the spectrum at the output of the FFCS when the FM noise signal of **Figure 4.2.2.2** applied to the RF input. The cancellation performance is close to 15 dB in a bandwidth of 5 MHz around the center frequency, and 25 dB in the frequency band between 2.5 and 8 MHz offset from pilot tone frequency.

Figure 4.3.4.2 – Spectral Measurement at the Output of the FFCS Trained with a 0 dBm Pilot Tone with FM Noise Signal Applied at the RF Input

Comparing the cancellation performance of **Figure 4.3.4.2** with **Figure 4.2.2.4**, it can be observed that the closed loop cancellation performance of the FFCS is not as good as can

be achieved by manual tuning. The cancellation level outside the 5 MHz band near center frequency is 13 dB less in closed loop operation of the FFCS compared with the cancellation level achieved by manual adjustment of the vector modulator. The cancellation level is further degraded in the 5 MHz band near the center frequency where the cancellation level is 25 dB less in closed loop operation of the FFCS compared with the cancellation level achieved by manual adjustment of the vector modulator.

The large difference in expected performance of closed loop operation of the FFCS is mostly due to the large loop bandwidth of the baseband processor which is in the order of a few hundred kHz. The noise components near the pilot tone frequency that compose the error signal contribute to a significant portion of the noise that is amplified and applied to the vector modulator. The modulation is then introduced in the cancellation signal. Since the noise in the cancellation signal is not correlated with the main path signal, it cannot be cancelled and is directly introduced in the main path. The closed loop performance of the FFCS can be improved by reducing the loop bandwidth.

Chapter 5 – Conclusion and Future Work

This chapter completes the thesis with a summary of the achieved objectives, and future recommendations for further areas of development and research. Finally, a discussion on potential technology applications of the FFCS closes the thesis.

Some significant technological issues that remain to be addressed to make this technology a viable complement to current duplexer technology have been identified. Further research in the area of feedforward noise cancellation is required to determine whether the current technological barriers can be overcome.

Although the FFCS may not immediately be a viable solution to complement a duplexer in the radio front-end of a cellular base station, it may have potential in other applications where the limitations of the current implementation are not significant.

5.1 – Summary of the FFCS Key Performance Metrics

This thesis developed and demonstrated the concept of feedforward noise cancellation applied in the context of a tunable notch filter operating at RF frequencies around 800 MHz. Implemented with current technology, the FFCS serves as the most basic and low complexity framework to design and build a practical actively tunable bandstop filter that works on the principle of feedforward noise cancellation. The FFCS developed has the potential application of complementing an RF front-end base station duplexer without any major overhaul.

The FFCS prototype achieves a peak cancellation level of 56 dB down from its minimum insertion loss of 2.77 dB. A cancellation level better than 50 dB was demonstrated over a bandwidth of 4.5 MHz, 40 dB over a bandwidth of 14.4 MHz, and 30 dB over a bandwidth of 45 MHz. The cancellation performance was maintained over a tuning range of 30 MHz around the design center frequency of 836.5 MHz. The tuning range of

the FFCS exceeds the design bandwidth of the FFCS with the performance degrading as the center frequency is tuned at frequencies further than the design frequency range. The FFCS is frequency agile. The FFCS can be tuned to a specific frequency band and achieve a maximum level of cancellation with the optimization done by an adaptive control system that uses a pilot tone as the center frequency reference.

The maximum cancellation bandwidth is determined by the adaptive control system's loop gain, noise rejection and sensitivity of the I/Q demodulator to detect very low power signals. Noise rejection is a function of loop bandwidth. The smaller the loop bandwidth, better the noise rejection. The tuning rate, also a function of loop bandwidth, becomes slower with smaller loop bandwidth. The loop bandwidth is ultimately determined by the loop filter. The cancellation bandwidth performance is limited by the delay mismatch in the main path. The tuning range is limited by the delay mismatch in the reference path to the I/Q demodulator.

A few more technological issues still need to be addressed in the implementation of the FFCS to make it a suitable complement solution to the duplexer problem. As it stands, the FFCS has no means of selectively excising noise components from the RF input frequency spectrum. It cancels frequencies across the entire input signal bandwidth because of the broadband cancellation "V" characteristic. Furthermore, the FFCS cannot handle a high power signal in the main path. Any signal in the main path enters the cancellation path. The circuitry of the cancellation path is designed to cope with signals below 0 dBm. High power signals upwards of +20 dBm entering the cancellation path may saturate or damage the sensitive circuitry.

5.2 – Future Modifications for Optimization

This thesis concentrates on the fundamental factors affecting cancellation performance, and practical technological issues specific to the chosen implementation. However, the reader should take note that other factors limit the achievable cancellation performance.

Many implementation and technological issues have been identified as a result of this research that could serve as potential research areas to improve the performance of the implemented feedforward noise cancellation system. Further research could potentially enable the feedforward technology to fit new applications.

5.2.1 – Required Improvements on the FFCS for Feasible Duplexer Complement

The main requirement to make the FFCS operational in a useful application is to introduce a receive band-pass or transmit band-stop filter after the sampling coupler in the cancellation path. The FFCS must selectively sample a very weak (-23 dBm) undesired signal adjacent in frequency (separated by a fraction bandwidth of the operating frequency) to a very strong (+45 dBm) desired signal. A receive band-pass or transmit band-stop would assure that the transmit signal does not get disturbed by the cancellation process. If a receive band-pass filter is used, the receive band noise only would be sampled. If a transmit band-stop filter is used, all the noise components around the transmit band would be sampled. Only the sampled spectral components can serve as cancellation signal. The same filter should be introduced after the error signal sampling coupler in the error signal path to protect the I/Q demodulator. The I/Q demodulator is not designed for input signals above +10 dBm. Furthermore, high power signal components in the range of + 10 dBm would desensitize the RF input of the I/Q demodulator to sense very low error signal components in the range of -80 dBm.

The immediate impact of inserting high rejection band-stop filters in the cancellation signal path and the error signal path is the requirement for longer delay lines. Increasing the delay increases the insertion loss in the main path.

Inserting a high rejection receive band-pass filter in the cancellation signal path and error signal path would also limit the tuning range and bandwidth of operation of the FFCS. In an application where wide tuning range is required, it would be preferable to use a transmit band-stop filter to conserve the frequency agility feature of the FFCS.

The choice of filter technology used to reject the transmit signal may have an impact on the dynamic range of the FFCS, essentially limiting the achievable cancellation performance. A SAW filter that can handle maximum power levels of +20 dBm would require 25 dB sampling directional couplers to prevent failure of the SAW filter. The error signal would be reduced by 15 dB. The sensitivity of the I/Q demodulator used drops sharply for RF signals below -65 dBm. An RF gain block would be required in the error signal path in presence of a 25 dB directional error sampling coupler, to assure the same cancellation performance as obtained with a 10 dB error sampling coupler. Ideally, the transmit bandstop filter would have a rejection at least as large as the SNR out of the power amplifier. If the rejection of the transmit bandstop filter is less, the signal will be re-introduced in the transmit path through the cancellation coupler and increase the noise floor in the transmit band, degrading the SNR of the transmitted signal.

In summary, instead of noise sampling coupler **CP2**, a frequency selective sampling coupler is required that would provide the possibility to sample a small signal (-26 dBm) in proximity (20% bandwidth separation around 850 MHz) of a large signal (+45 dBm). The frequency selective sampling coupler would introduce minimal insertion loss (less than 0.15 dB) to the desired signal in the transmit band, while keeping low amplitude and phase ripple to the sampled noise signal during sampling. The frequency selective sampling coupler would provide a rejection to transmit band signal at the coupled port as high as the SNR out of the PA (typically -40 to -60 dBc) to the transmit band.

5.2.2 – Other Cancellation Performance Factors

Slight improvements in cancellation performance can be achieved by considering second order effects. Shielding is important to reduce external interference effects. Proper matching of the different subcircuits is important to avoid reflections between the different subcircuits. Signal reflections at the interface between two subcircuits in the main path or cancellation path would contribute to reducing the correlation between the

main signal and cancellation signal. Analysis of shielding and return loss between subcircuits and limited isolation in power dividers, signal combiners and directional couplers have been reported to affect cancellation performance [38, 53].

5.2.3 - Further Theoretical Aspects that Should be Understood to Improve the Design of the FFCS

It would be interesting to establish the correlation function between the main signal and the cancellation signal to analyze noise performance and establish practical cancellation performance limits. Initial analyses found in [49].

The noise figure of the vector modulator should be determined.

5.2.4 – RF Architecture Optimizations

Use of phase equalizers: Some modifications could be implemented to the current architecture to improve the cancellation or control system performance. In [37], phase equalizers were used instead of delay lines. Phase equalizers have the inverse non-linear dependence of phase on frequency that the gain blocks or other components would have. The phase equalizer essentially matches (equalizes) the group delay characteristics of the main path and cancellation path to improve broadband performance.

Use of other delay line technology: The delay lines in the FFCS are implemented using a significant length of coaxial cables. Other technological options are also available to delay an RF signal. Cavity and surface-mount delay line filters are available [54]. Potential benefits of using Cavity and surface mount delay line filters would be reduced size, lower insertion loss and better matching. A cavity filter can deal with high power signals present in the main path, while a surface mount delay line would be suitable for low power signals present in the pilot tone reference path. A further improvement would be to fine-tune the time delays in the pilot tone reference path using active components

[30]. Adaptively fine tuning the delay would avoid lengthy manual trimming of the delay line coaxial cables. Improving the delay in the pilot tone reference path would improve the cancellation performance level, optimization convergence time. Optimizing the delay in the main path could provide the ability to perform dynamic bandwidth adjustments. Optimizing the delay in the reference path could provide the ability to adapt the system to the frequency band of operation by maintaining the vector modulator and I/Q demodulator aligned along the same reference plane extending, extending the frequency tuning range.

Use MEMS switches to change hardware configuration: The use of MEMS could offer the possibility to have a multi-mode system that could operate in different frequency bands. For example RF MEMS switch in the design could allow to switch between delay lines required for a system operated in either the 800 MHz band or the 2 GHz band. Although some components may be wideband such as the gain blocks and the directional couplers, other components such as the vector modulator may be more frequency selective. The use of MEMS could also allow to switch between two function blocks without requiring two different architectures.

Reduce main path insertion loss: One of the most important practical performance characteristic to control is the insertion loss in the main path. The insertion loss depends on the main path delay line length and on the coupling values of the signal injection and signal sampling directional couplers in the main path. By optimizing the coupling values of the directional couplers present in the main path, it could be possible to reduce the insertion loss by sacrificing some cancellation performance. Quantifying the change in cancellation performance as a function of insertion loss and coupler optimization would be useful to specify the performance specifications of the system in a particular application. Using dielectric materials with less loss and minimizing the design by integrating the system into a chip solution could bring an increase in performance and decrease the insertion loss significantly. Porting the FFCS design to a chip size solution

could also reduce the cost and offer better tolerance control and more precise delay matching.

Use different pilot tone control strategy: In a sense, injecting a pilot tone signal in the main path is a little counter productive as a noise reduction technique. The injected pilot tone must be above the noise level to be reduced in order to be detected and is only removed up to the level of cancellation achieved. Instead of a single tone pilot tone, a spread spectrum signal could be injected in the main path as the correlation reference signal. The spread spectrum reference signal could be injected at the same level or lower than the noise level to be reduced, introducing negligible noise in the main path. Depending on the communication standard used, and provided good control system predictability, the pilot tone could be injected in guard bands or at an offset of the receive channel to minimize the added noise in the receive band.

The noise in the baseband section could be reduced by using some other detection technique for the error signal. A multiple tone pilot control signal or spread spectrum techniques could be used to achieve noise immunity without sacrificing dynamic tracking performance. Because the power level of the noise signal to be cancelled is near the level of the pilot tone, the error estimate becomes a composite of the pilot tone and residual noise. The accuracy of the error signal is hence reduced and becomes noisy. The only way to reduce the noisy spurs is to reduce the loop bandwidth and gain of the control system.

Reduce quadrature phase errors in the vector modulator: The quadrature error between the I and Q-branch of the vector modulator could be reduced by using balanced bi-phase modulators [55]. An adaptive control system could be used around the vector modulator to assure minimal quadrature error between the I and Q-branch.

Use I/Q demodulator with higher sensitivity: The I/Q demodulator has limited sensitivity which in turn limits the maximum cancellation level achievable. The effect of

limited demodulator sensitivity may also manifest in similarity to the effect of DC offsets in the baseband section, and may be minimized by increasing the control loop gain. Demodulator sensitivity may lead to the degree of offset of the point of maximum cancellation from the pilot tone frequency and is also a function of delay match. The better the delay match, the wider the cancellation bandwidth. Hence, the point of maximum cancellation may be further in frequency from a given point of cancellation along either sloping side of the “V” characteristic which the demodulator is able to detect. A gain block could also be introduced in the error path to obtain the same effect.

5.2.5 – Baseband Processor Optimizations

The noise and DC offsets in the baseband section are the primary cause of reduced cancellation performance during closed loop operation. DC offsets cause incorrect convergence of the cancellation performance, and introduce a static error limited in part by the loop gain. Any noise added in the baseband section is directly communicated to the PIN diodes and add noise to the cancellation signal in the vector modulator. The noise figure of the vector modulator could be reduced by reducing the noise in the baseband section. It is expected that optimizing noise figure and controlling the DC offsets in the baseband processor would yield a significant reduction in the static error, increasing cancellation performance and increase the bandwidth performance of the FFCS.

Reduce DC offsets: DC offsets could be monitored and some compensation techniques could be used to nullify the negative effect of the DC offset [33]. The use of chopping signals could allow to monitor the DC offsets and calibrate the system while it is in operation. Another benefits of having an adaptive DC offset nulling technique would be avoiding the need to re-tune the circuit in changing environmental conditions.

Use intermediate frequency (IF): The detection technique used to decompose the error signal has some technical drawbacks. The noise level at the baseband output of the I/Q

demodulator is relatively high at frequency components very near DC. The high noise level is caused by the 1/frequency flicker noise characteristic inherent in all natural phenomenon. The noise level at the baseband output (especially near DC) restricts the sensitivity of the I/Q demodulator. Band-limiting and averaging techniques are ineffective at improving the accuracy of decomposed baseband near-DC component because of the nature of flicker noise. Down-converting the error signal at an intermediate frequency (IF) for signal processing is common practice to avoid the high flicker noise problem at low frequencies. In [34], the signals are demodulated to an intermediate frequency (IF) for processing to avoid the DC offsets from mixers.

5.2.6 – Outstanding Experimental Observations Requiring Justification

Loss of lock was observed when the main path power is too high. Also, if not already locked, the system cannot lock if the main path power level is already at a certain level below the power level above which the FFCS loses lock. It is suspected that the gain blocks or the PIN diodes in the cancellation path operate with high non-linearity or become saturated. The noise level in the cancellation path is raised up to the level of the signal in the main path, essentially masking the pilot tone signal. Therefore, the control system is unable to produce a proper control signal.

5.2.7 – Potential Topics of Future In-Depth Research

The FFCS should be tested with receive band-pass or transmit band-stop filter in the cancellation and error paths. The FFCS should also be tested with a high power input signal in presence of low power noise.

Determine the minimum possible delay through the cancellation path of an optimized architecture, find the practical minimum possible insertion loss and the smallest possible implementation for the FFCS with current IC technologies given the range of power levels it would have to deal with (+45 dBm).

5.3 – Technology Potential

The pilot tone for the prototype design is to be supplied via an external RF signal generator. In a product implementation of the circuit the pilot tone will consist of a phase locked VCO. Phase locking of the pilot tone will enable digital control of the pilot tone. The pilot tone in a base-station application would be slaved to the base station local oscillator synthesizer digital control line and phase locked to the base-station reference oscillator. Slaving the pilot tone to the base-station local oscillator in this manner will enable the noise cancellation system center frequency to track the receive frequency even in the event of the base-station switching to a new receive channel.

In another scenario, the pilot tone synthesizer can be locked to a received interference signal in order that the cancellation notch can track the interferer (provided the rate of change in frequency of the interferer does not exceed the control loop bandwidth which may be limited to a fraction of the noise cancellation bandwidth). The interference tracking and cancellation application would also find use in the receive path of the radio system.

Two independently tunable FFCS loops could be cascaded or paralleled to achieve wider bandwidth or other cancellation characteristics as demonstrated in [42] for a feedforward amplifier. Investigating the options of cascaded stages further could bring some interesting filtering characteristic to the RF field. For example, a filter characteristic could be implemented that has multiple independently tunable notches in the receive band. The multiple unit aspect of the design was not investigated in this thesis.

The feedforward cancellation technology has great potential to realize tunable band-pass filters with wide tunable range and good bandwidth performance. Further research in the field could help to find new applications for feedforward cancellation, better performance and characterize the technology to help engineers in the design process.

The FFCS could have potential use in system identification applications. For example, the main path delay line could be replaced by a device under test (DUT) to measure the electrical delay, phase shift and amplitude characteristics. The π -attenuator in the cancellation path could then be replaced by a variable gain amplifier to automatically adjust the insertion loss of the cancellation path to compensate for an increase or decrease in insertion loss in the main path due to the change in delay or gain of the component to characterize.

The signal modulation independent performance makes the FFCS attractive for use with many types of communication systems and defense applications. In applications where lower signal levels are used, feedforward cancellation technology yields itself well to miniaturization at frequencies above 800 MHz. The concept of the FFCS could also be adapted to feedforward amplifiers to make them frequency agile.

The FFCS could have potential use as a defense against jammers. The FFCS would lock on and track an interference signal, whether it be a continuous wave (CW) tone jammer or a frequency hopping jammer, and cancel it. In such an application, the pilot signal would be the jamming signal.

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APPENDIX A - Derivation of Cancellation Performance Equations

A1 - Cancellation as a Function of Amplitude Imbalance Only

a) Substituting (2.1) and (2.3) in (2.7) yields (A.1)

$$v_{err}(t) = v_{main}(t) + v_{can}(t)$$

$$v_{err}(t) = V_{main} \cdot \cos(2\pi \cdot f_c \cdot t - \beta_c \cdot l_{main} + \phi_{main}) \dots \\ + (V_{main} + V_{diff}) \cdot \cos[2\pi \cdot f_c \cdot t - \beta_c \cdot (l_{main} + \Delta l) + 180^\circ + (\phi_{main} + \theta_{err})]$$

$$v_{err}(t) = V_{main} \cdot \cos(2\pi \cdot f_c \cdot t - \beta_c \cdot l_{main} + \phi_{main}) \dots \\ + -(V_{main} + V_{diff}) \cdot \cos[[2\pi \cdot f_c \cdot t - \beta_c \cdot (l_{main} + \Delta l)] + (\phi_{main} + \theta_{err})] \quad (A.1)$$

b) Simplifying for the case where: $\theta_{err} = 0$ $\Delta l = 0$

$$v_{err}(t) = -V_{diff} \cdot \cos(2\pi \cdot f_c \cdot t - \beta_c \cdot l_{main} + \phi_{main}) \quad (A.2)$$

c) Substituting (2.1) in (2.8) and simplifying, yields (A.4)

$$P_{main} = \frac{1}{T_c} \int_0^{T_c} \frac{v_{main}(t)^2}{Z_o} dt = \frac{1}{T_c} \int_0^{T_c} \frac{(V_{main} \cdot \cos(2\pi \cdot f_c \cdot t - \beta_c \cdot l_{main} + \phi_{main}))^2}{Z_o} dt \\ P_{main} = \frac{V_{main}^2}{T_c \cdot Z_o} \int_0^{T_c} \cos^2(2\pi \cdot f_c \cdot t - \beta_c \cdot l_{main} + \phi_{main})^2 dt \quad (A.3)$$

Using trigonometric identity: $\cos^2(x) = \frac{1 + \cos(2x)}{2}$

$$P_{main} = \frac{V_{main}^2}{T_c \cdot Z_o} \int_0^{T_c} \left[\frac{1}{2} + \frac{1}{2} \cdot \cos[2 \cdot (2\pi \cdot f_c \cdot t - \beta_c \cdot l_{main} + \phi_{main})] \right] dt$$

$$P_{main} = \frac{V_{main}^2}{2 T_c \cdot Z_o} \int_0^{T_c} (1) dt + \frac{1}{2} \cdot \int_0^{T_c} \cos(4\pi \cdot f_c \cdot t - 2\beta_c \cdot l_{main} + 2\phi_{main}) dt$$

Using integral identity: $\int \cos(u) du = \sin(u)$

u $du = 4\pi \cdot f \cdot dt$

$$P_{main} = \frac{V_{main}^2}{2 T_c \cdot Z_o} \int_0^{T_c} (1) dt + \frac{1}{2} \cdot \int_0^{T_c} \cos(4\pi f_c t - 2\beta_c \cdot l_{main} + 2\phi_{main}) \cdot \left(\frac{4\pi \cdot f \cdot dt}{4\pi \cdot f \cdot dt} \right) dt$$

$$P_{\text{main}} = \frac{V_{\text{main}}^2}{2 \cdot T_c \cdot Z_o} \cdot (T_c - 0) + \frac{1}{2} \cdot \frac{\sin[4 \cdot \pi \cdot f_c \cdot (T_c) - 2 \cdot \beta_c \cdot l_{\text{main}} + 2 \cdot \phi_{\text{main}}]}{4 \cdot \pi \cdot f_c} \dots$$

$$+ - \frac{\sin[4 \cdot \pi \cdot f_c \cdot (0) - 2 \cdot \beta_c \cdot l_{\text{main}} + 2 \cdot \phi_{\text{main}}]}{4 \cdot \pi \cdot f_c}$$

Since:

$$\frac{1}{2} \cdot \frac{\sin[4 \cdot \pi \cdot f_c \cdot (T_c) - 2 \cdot \beta_c \cdot l_{\text{main}} + 2 \cdot \phi_{\text{main}}]}{4 \cdot \pi \cdot f_c} = \frac{\sin[4 \cdot \pi \cdot f_c \cdot (0) - 2 \cdot \beta_c \cdot l_{\text{main}} + 2 \cdot \phi_{\text{main}}]}{4 \cdot \pi \cdot f_c}$$

$$P_{\text{main}} = \frac{V_{\text{main}}^2}{2 \cdot Z_o} \quad (\text{A.4})$$

d) Substituting (A.1) in (2.9) and simplifying, yields (A.6)

$$P_{\text{err}} = \frac{1}{T_c} \cdot \int_0^{T_c} \frac{v_{\text{err}}(t)^2}{Z_o} dt$$

$$P_{\text{err}} = \frac{1}{T_c} \cdot \int_0^{T_c} \frac{v_{\text{err}}(t)^2}{Z_o} dt = \frac{1}{T_c} \cdot \int_0^{T_c} \frac{(-V_{\text{diff}} \cdot \cos(2 \cdot \pi \cdot f_c \cdot t - \beta_c \cdot l_{\text{main}} + \phi_{\text{main}}))^2}{Z_o} dt$$

$$P_{\text{err}} = \frac{V_{\text{diff}}^2}{T_c \cdot Z_o} \cdot \int_0^{T_c} \cos^2(2 \cdot \pi \cdot f_c \cdot t - \beta_c \cdot l_{\text{main}} + \phi_{\text{main}}) dt \quad (\text{A.5})$$

Noticing that (A.5) has the same structure as (A.3) from part (c):

$$P_{\text{err}} = \frac{V_{\text{diff}}^2}{2 \cdot Z_o} \quad (\text{A.6})$$

Substituting (A.4) and (A.6) in (2.10) and simplifying yields (A.7)

$$CP = 10 \log \left(\frac{P_{\text{err}}}{P_{\text{main}}} \right) = 10 \cdot \log \left(\frac{V_{\text{diff}}^2}{2 \cdot Z_o} \cdot \frac{2 \cdot Z_o}{V_{\text{main}}^2} \right)$$

$$CP = 10 \log \left(\frac{V_{\text{diff}}^2}{V_{\text{main}}^2} \right) \quad (\text{A.7})$$

e) It is more convenient to express the cancellation performance as a function of normalized amplitude imbalance (expressed in dB).

Figure A.1 - Cancellation by Vector Addition

Figure A.1 a), shows the main signal and the cancellation signal on a polar plot. The phase error is shown as θ_{err} and the amplitude imbalance is shown as α_{dB} . In **Figure A.1 b)**, V_{can} is vectorially added to V_{main} to produce the V_{err} . The signals are normalized in magnitude to V_{main} .

Using the cosine rule: $A^2 = B^2 + C^2 - 2 \cdot B \cdot C \cdot \cos(\theta_A)$

The error vector can be found geometrically:

$$\left(\frac{V_{err}}{V_{main}} \right)^2 = \left(\frac{V_{main}}{V_{main}} \right)^2 + \left(\frac{V_{main} + V_{diff}}{V_{main}} \right)^2 - 2 \cdot \frac{V_{main}}{V_{main}} \cdot \frac{V_{main} + V_{diff}}{V_{main}} \cdot \cos(\theta_{err})$$

f) Simplifying and knowing that $\theta_{err} = 0$ yields (A.8)

$$\left(\frac{V_{err}}{V_{main}} \right)^2 = 1 + \left(\frac{V_{main} + V_{diff}}{V_{main}} \right)^2 - 2 \cdot \left(\frac{V_{main} + V_{diff}}{V_{main}} \right) \quad (A.8)$$

Reducing the right side (A.8) yields:

$$\begin{aligned} \left(\frac{V_{err}}{V_{main}} \right)^2 &= 1 + \frac{V_{main}^2}{V_{main}^2} + \frac{2 \cdot V_{main} \cdot V_{diff}}{V_{main}^2} + \frac{V_{diff}^2}{V_{main}^2} - 2 \cdot \frac{V_{main} + V_{diff}}{V_{main}} \\ \left(\frac{V_{err}}{V_{main}} \right)^2 &= 1 + 1 + \frac{2 \cdot V_{diff}}{V_{main}} + \frac{V_{diff}^2}{V_{main}^2} - 2 - 2 \cdot \frac{V_{diff}}{V_{main}} = \frac{V_{diff}^2}{V_{main}^2} \end{aligned} \quad (A.9)$$

As expected, (A.9) is the same ratio in (A.7). Using the form of (A.8), (A.7) can be expressed as a function of amplitude imbalance to obtain equation (2.1.1.3). Deriving the cancellation performance by trigonometry can be easier and yield good insight in the phenomenon than deriving the analytical expression by integrating the instantaneous equations.

Note that: $\alpha = \frac{V_{\text{main}} + V_{\text{diff}}}{V_{\text{main}}}$: Normalized amplitude imbalance

$$\text{CP} = 10 \cdot \log(1 - 2 \cdot \alpha + \alpha^2)$$

The analytical expression of the cancellation performance as a function of amplitude imbalance when expressed in (dB) becomes:

$$\text{CP}(\alpha_{\text{dB}}) = 10 \cdot \log\left(1 - 2 \cdot 10^{\frac{\alpha_{\text{dB}}}{20}} + 10^{\frac{2 \alpha_{\text{dB}}}{20}}\right) \quad ** (2.1.1.3)$$

A2 - Cancellation as a Function of Phase Error Only

Using the trigonometric approach of A1 e) and f), the normalized error vector can be simplified and derived knowing:

$$\alpha_{\text{dB}} := 0 \text{ dB} \quad \text{i.e.:} \quad |V_{\text{main}}| = |V_{\text{can}}| \quad \text{i.e.:} \quad V_{\text{diff}} = 0$$

$$\left(\frac{V_{\text{err}}}{V_{\text{main}}}\right)^2 = 1^2 + 1^2 - 2 \cdot \cos(\theta_{\text{err}}) = 2 - 2 \cdot \cos(\theta_{\text{err}})$$

The cancellation performance as a function of phase error only can therefore be expressed as equation (2.1.2.1).

$$\text{CP}(\theta_{\text{err}}) = 10 \cdot \log[2 \cdot (1 - \cos(\theta_{\text{err}}))] \quad ** (2.1.2.1)$$

A3 - Cancellation as a Function of Amplitude Imbalance and Phase Error

Using the trigonometric approach of A1 e) and f), the normalized error vector can be simplified and derived:

$$\left(\frac{V_{\text{err}}}{V_{\text{main}}}\right)^2 = 1 + \left(\frac{V_{\text{main}} + V_{\text{diff}}}{V_{\text{main}}}\right)^2 - 2 \cdot \left(\frac{V_{\text{main}} + V_{\text{diff}}}{V_{\text{main}}}\right) \cdot \cos(\theta_{\text{err}})$$

The cancellation performance as a function of amplitude imbalance (expressed in dB) and phase error (in °) can therefore be expressed as equation (2.1.3.1).

$$\text{CP}(\alpha_{\text{dB}}, \theta_{\text{err}}) = 10 \cdot \log\left(1 - 2 \cdot 10^{\frac{\alpha_{\text{dB}}}{20}} \cdot \cos(\theta_{\text{err}}) + 10^{\frac{2 \alpha_{\text{dB}}}{20}}\right) \quad ** (2.1.3.1)$$

A4 - Cancellation as a Function of Delay Mismatch Only

a) The relative phase shift between each path at the point of recombination can be expressed as a function of frequency assuming perfect phase match at the design frequency:

The main path and cancellation path phase shift as a function of frequency by combining (2.1.4.1) and (2.1.4.2) into (2.1.4.3):

$$\theta_{\text{main}}(f) = \frac{-2 \cdot \pi \cdot l_{\text{main}} \cdot \sqrt{\epsilon_r}}{c} \cdot f \quad \theta_{\text{can}}(f) = \frac{-2 \cdot \pi \cdot l_{\text{can}} \cdot \sqrt{\epsilon_r}}{c} \cdot f$$

The main path and cancellation path phase shift at the center frequency f_c is:

$$\theta_{\text{main}_c} = \frac{-2 \cdot \pi \cdot l_{\text{main}} \cdot \sqrt{\epsilon_r}}{c} \cdot f_c \quad \theta_{\text{can}_c} = \frac{-2 \cdot \pi \cdot l_{\text{can}} \cdot \sqrt{\epsilon_r}}{c} \cdot f_c$$

The main path phase shift is referenced to zero phase by subtracting the center frequency phase shift:

$$\theta_{\text{main}_o}(f) = \theta_{\text{main}}(f) - \theta_{\text{main}_c}$$

$$\phi_{\text{main}_o} = \frac{-2 \cdot \pi \cdot l_{\text{main}} \cdot \sqrt{\epsilon_r}}{c} \cdot f - \frac{-2 \cdot \pi \cdot l_{\text{main}} \cdot \sqrt{\epsilon_r}}{c} \cdot f_c = \frac{-2 \cdot \pi \cdot \sqrt{\epsilon_r}}{c} \cdot l_{\text{main}} \cdot (f - f_c)$$

The cancellation path phase shift is referenced to zero phase by subtracting the center frequency phase shift:

$$\theta_{\text{can}_o}(f) = \theta_{\text{can}}(f) - \theta_{\text{can}_c}$$

$$\theta_{\text{can}_o}(f) = \frac{-2 \cdot \pi \cdot l_{\text{can}} \cdot \sqrt{\epsilon_r}}{c} \cdot f - \frac{-2 \cdot \pi \cdot l_{\text{can}} \cdot \sqrt{\epsilon_r}}{c} \cdot f_c = \frac{-2 \cdot \pi \cdot \sqrt{\epsilon_r}}{c} \cdot l_{\text{can}} \cdot (f - f_c)$$

The phase difference is the same as defined previously:

$$\Delta\theta = \theta_{\text{can}_o}(f) - \theta_{\text{main}_o}(f) = \left(\frac{-2 \cdot \pi \cdot \sqrt{\epsilon_r}}{c} \right) \cdot (f - f_c) \cdot (l_{\text{can}} - l_{\text{main}}) \quad (\text{A.10})$$

Multiplying (A.10) by (f_c/f_c) and reorganizing:

$$\Delta\theta = \left[\frac{2 \cdot \pi}{\left(\frac{c}{\sqrt{\epsilon_r} \cdot f_c} \right)} \right] \cdot \left(\frac{f}{f_c} - \frac{f_c}{f_c} \right) \cdot \Delta l \quad \text{Knowing: } \frac{2 \cdot \pi}{\left(\frac{c}{\sqrt{\epsilon_r} \cdot f_c} \right)} \text{ is: } \beta_c = \frac{2 \cdot \pi}{\lambda_c}$$

Note: λ_c is the wavelength at the center frequency.

$$\Delta\theta = \Delta l \cdot (\beta_c - \beta(f)) = (2 \cdot \pi \cdot \Delta l) \cdot \left(\frac{1}{\lambda_c} - \frac{1}{\lambda_{\text{eff}}(f)} \right) = \frac{2 \cdot \pi \cdot \Delta l}{\lambda_c} \cdot \left(1 - \frac{\lambda_c}{\lambda_{\text{eff}}(f)} \right) = \beta_c \cdot \Delta l \cdot \left(1 - \frac{f}{f_c} \right)$$

The difference in wavelength expressed in meters (m) at the design frequency is: Δl

The difference in the *number of wavelength* at the design frequency is: $\lambda_{\text{err}} = \frac{\Delta l}{\lambda_c}$ (A.11)

The relative phase shift between the two paths resulting from delay mismatch is given by:

$$\Delta\theta(f) = \beta_c \cdot \Delta l \cdot \left(1 - \frac{f}{f_c} \right) \quad (\text{A.12})$$

In (A.12), the product $\beta_c \cdot \Delta l$ is the angular difference in wavelengths at the center frequency. (A.12) indicates that the angular difference in wavelength is proportional to the frequency.

The relative phase shift as a function of frequency is also expressed as a function of wavelength difference at the design frequency by combining (A.11) and (A.12) to yield (2.1.4.5).

$$\Delta\theta\left(\frac{f}{f_c}, \lambda_{\text{err}}\right) = 2 \cdot \pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c} \right) \quad \text{**(2.1.4.5)}$$

b) (2.1.4.5) is then substituted for θ_{err} in (2.14) to express the cancellation performance as a function of delay mismatch in terms of the difference in the number of wavelength between the two paths at the point of cancellation, resulting in (2.1.4.6).

$$\text{CP}\left(\frac{f}{f_c}, \lambda_{\text{err}}\right) = 10 \cdot \log \left[2 \cdot \left[1 - \cos \left[2 \cdot \pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c} \right) \right] \right] \right] \quad \text{**(2.1.4.6)}$$

A5 - Cancellation as a Function of All Fundamental Performance Determining Factors

a) The cancellation performance as a function of all fundamental performance determining factors is derived by adding (2.1.4.5) to θ_{err} in (2.15).

**(2.1.5.1)

$$\text{CP}\left(\alpha_{\text{dB}}, \theta_{\text{err}}, \lambda_{\text{err}}, \frac{f}{f_c}\right) = 10 \cdot \log \left[1 - 2 \cdot 10^{\frac{\alpha_{\text{dB}}}{20}} \cdot \cos \left[2 \cdot \pi \cdot \lambda_{\text{err}} \cdot \left(1 - \frac{f}{f_c} \right) + \theta_{\text{err}} \right] + 10^{\frac{2 \alpha_{\text{dB}}}{20}} \right]$$

b) To help gain a better understanding the simultaneous effects of all the factors affecting the cancellation performance, two surface plots of (2.1.5.1) are shown in **Figure 2.1.5.1** for $\lambda_{\text{err}} = 1/3$. The top plot corresponds to $\alpha_{\text{dB}} = 0.1$ dB and the bottom plot corresponds to $\alpha_{\text{dB}} = 0$.

It can be observed from Figure 2.1.5.1 that the contour levels of the cancellation performance surface form straight lines in the X-Y plane defined by the phase error (X) and normalized amplitude imbalance (Y). The slope of the line can be found.

Maximum cancellation performance is achieved when: $\alpha_{dB} := 0 \text{ dB}$

$$CP = 10 \cdot \log[1 + (1) - 2 \cdot (1) \cdot \cos[2 \pi \cdot \lambda_{err} \cdot (1 - f_n) + \theta_{err}]] = -\infty$$

$$1 + (1) - 2 \cdot (1) \cdot \cos[2 \pi \cdot \lambda_{err} \cdot (1 - f_n) + \theta_{err}] = 0$$

$$2 - 2 \cdot \cos[2 \pi \cdot \lambda_{err} \cdot (1 - f_n) + \theta_{err}] = 0$$

$$\cos[2 \pi \cdot \lambda_{err} \cdot (1 - f_n) + \theta_{err}] = 1$$

$$2 \pi \cdot \lambda_{err} \cdot (1 - f_n) + \theta_{err} = \arccos(1) = 0$$

$$f_n = -\left(\frac{-\theta_{err}}{2 \pi \cdot \lambda_{err}} - 1\right)$$

Plotting the line of f_n as a function of θ_{err} for: $cp := -\infty \text{ dB}$ $\lambda_{err} := \frac{1}{3}$

$$f_n(\theta_{err}) = \frac{\theta_{err}}{2 \cdot \pi \cdot \lambda_{err}} + 1 \quad \text{**(2.1.5.2)}$$

The slope of the line is:

$$m = \frac{1}{2 \cdot \pi \cdot \lambda_{err}} \quad \text{(A.13)}$$

A6 - I and Q Component Imbalance as a Function of Amplitude Imbalance and Phase Error

a) The I and Q component imbalance between the main signal and cancellation signal can be derived to give a more intuitive understanding of the cancellation performance with respect to the chosen I/Q control scheme. I and Q components imbalance can be derived directly from the Amplitude imbalance and phase error between the main signal and cancellation signal.

Given the main and cancellation signals:

$$v_{\text{main}}(t) = V_{\text{main}} \cdot \cos(2 \cdot \pi \cdot f_c \cdot t + \phi_{\text{main}})$$

$$v_{\text{can}}(t) = V_{\text{main}} \cdot 10^{\frac{\alpha_{\text{dB}}}{20}} \cdot \cos(2 \cdot \pi \cdot f_c \cdot t + 180^\circ + \phi_{\text{main}} + \theta_{\text{err}})$$

and assuming: $V_{\text{main}} = 1$ and $\phi_{\text{main}} = 0^\circ$

The main and cancellation signals are plotted in the I/Q cartesian plane of **Figure 2.2.3.2 a)** for a given phase error θ_{err} and amplitude imbalance α_{dB} .

Figure A.3 - Cancellation Residual Error I/Q Vectors (Normalized)

The I/Q reference plane is first defined by taking the Main vector as a reference. The I/Q plane is defined as the standard cartesian coordinates where the X-axis corresponds to the I component and the Y-axis corresponds to the Q component in **Figure 2.2.3.2**. The I and Q component imbalance is taken with respect to the Main reference vector. In the following, the reference vector is taken as $v_{\text{main}}(t)$ at $t = 0$, with $\phi_{\text{main}} = 0^\circ$, and ignoring the phase constant delays:

b) In terms of I and Q components, the main signal is expressed as equations (A.14) and (A.15) respectively.

$$I_{\text{main}} = V_{\text{main}} \cdot \cos(\phi_{\text{main}}) \quad (\text{A.14})$$

$$Q_{\text{main}} = V_{\text{main}} \cdot \sin(\phi_{\text{main}}) \quad (\text{A.15})$$

c) In terms of I and Q components, the cancellation signal is expressed as equations (A.16) and (A.17) respectively.

$$I_{\text{can}} = V_{\text{main}} \cdot \left(10^{\frac{\alpha_{\text{dB}}}{20}}\right) \cdot \cos(180 \text{ deg} + \phi_{\text{main}} + \theta_{\text{err}}) \quad (\text{A.16})$$

$$Q_{\text{can}} = V_{\text{main}} \cdot \left(10^{\frac{\alpha_{\text{dB}}}{20}}\right) \cdot \sin(180 \text{ deg} + \phi_{\text{main}} + \theta_{\text{err}}) \quad (\text{A.17})$$

The vector component resulting from the vector addition of the I and Q components of the main signal with the cancellation signal is the cancellation residual and corresponds to the $(\Delta I, \Delta Q)$ vector in **Figure A.3**. The error vector I and Q components are defined by equations (A.18) and (A.19).

$$\Delta I = I_{\text{main}} + I_{\text{can}} \quad (\text{A.18})$$

$$\Delta Q = Q_{\text{main}} + Q_{\text{can}} \quad (\text{A.19})$$

Normalizing and simplifying:

$$\Delta I = \cos(\phi_{\text{main}}) + \left(10^{\frac{\alpha_{\text{dB}}}{20}}\right) \cdot \cos(180 \text{ deg} + \phi_{\text{main}} + \theta_{\text{err}})$$

$$\Delta Q = \sin(\phi_{\text{main}}) + \left(10^{\frac{\alpha_{\text{dB}}}{20}}\right) \cdot \sin(180 \text{ deg} + \phi_{\text{main}} + \theta_{\text{err}})$$

$$\Delta I(\alpha_{\text{dB}}, \theta_{\text{err}}) = 1 - \left(10^{\frac{\alpha_{\text{dB}}}{20}}\right) \cdot \cos(\theta_{\text{err}}) \quad \text{**(2.2.3.1)}$$

$$\Delta Q(\alpha_{\text{dB}}, \theta_{\text{err}}) = - \left(10^{\frac{\alpha_{\text{dB}}}{20}}\right) \cdot \sin(\theta_{\text{err}}) \quad \text{**(2.2.3.2)}$$

d) Functions to convert amplitude imbalance and phase imbalance to an I component difference and Q component difference between the main signal and cancellation signal can be expressed as equations (2.2.3.3) and (2.2.3.4) respectively. Taking the negative of (2.2.3.1) and (2.2.3.2) yields the error estimate that serves as a noisy estimate of the gradient

$$\varepsilon_I(\alpha_{\text{dB}}, \theta_{\text{err}}) := 10^{\frac{\alpha_{\text{dB}}}{20}} \cdot \cos(\theta_{\text{err}}) + 1 \quad \text{**(2.2.3.3)}$$

$$\varepsilon_Q(\alpha_{\text{dB}}, \theta_{\text{err}}) := -10^{\frac{\alpha_{\text{dB}}}{20}} \cdot \sin(\theta_{\text{err}}) \quad \text{**(2.2.3.4)}$$

APPENDIX B - Sample Theoretical Calculations

B1 - Theoretical PCB Transmission Characteristics

a) PCB Substrate Parameters:

Substrate:	PARCNelco N4000-13
Relative Dielectric Constant:	$\epsilon_r := 3.9$
Substrate Thickness:	$H_{\text{sub}} := 12 \text{ mil}$
Relative Permeability:	$\mu_r := 1$
Upper Ground Plane to Substrate Spacing:	$3.9 \cdot 10^{-34} \text{ mil}$: Cover Height
Conductor Thickness:	$T_{\text{cnd}} := 0.984 \text{ mil}$ $T_{\text{cnd}} = 25 \text{ } \mu\text{m}$: 1 oz Copper
Conductor Conductivity:	$\text{Cond} := 4.1 \cdot 10^7 (\Omega \cdot \text{m})^{-1}$
Dielectric Loss Tangent:	$\text{TanD} := 0.014$
Conductor Surface Roughness:	$\text{Rough} := 0 \text{ mil}$
Conductor Material:	Copper with Emersion Silver Plating

b) Operating Conditions

Upper Frequency:	$f_h := 849 \text{ MHz}$
Center Frequency:	$f_c := 836.5 \text{ MHz}$
Lower Frequency:	$f_l := 824 \text{ MHz}$
Characteristic Impedance:	$Z_0 := 50 \Omega$

c) The wavelength as a function of frequency is defined as:

$$\lambda(f, \epsilon) := \frac{c}{f \sqrt{\epsilon}} \quad \text{Where: } c := 2.99792458 \cdot 10^8 \frac{\text{m}}{\text{s}} \quad (\text{Speed of light in free space})$$

The ideal wavelength at f_c :

$$\lambda_{fc} := \lambda(f_c, \epsilon_r) \quad \lambda_{fc} = 18.148 \text{ cm}$$

$$\lambda_{fc} = 7144.78 \text{ mil}$$

d) In microstrip lines, the effective propagation constant $\gamma = (\alpha + j\beta)$ may be calculated as follows:

Reference: S.R. Pennock & P.R. Shepherd, "Microwave Engineering With Wireless Applications," McGraw Hill, 1998

The w/h ratio for a given impedance and relative dielectric constant is given by:

$$w_{\text{'h}}(\epsilon_r, Z_0) := \frac{8 \cdot \sqrt{\left(\frac{Z_0}{e^{42.4 \Omega}} \sqrt{\epsilon_r + 1} - 1 \right) \cdot \left(\frac{7 + \frac{4}{\epsilon_r}}{11} \right) + \frac{1 + \frac{1}{\epsilon_r}}{0.81}}}{\frac{Z_0}{e^{42.4 \Omega}} \sqrt{\epsilon_r + 1} - 1} \quad w_{\text{'h}}(\epsilon_r, Z_0) = 2.081$$

The width of the transmission line is:

$$w_{\text{txline}} := w_{\text{'h}}(\epsilon_r, Z_0) \cdot H_{\text{sub}} \quad w_{\text{txline}} = 0.634 \text{ °mm}$$

$$w_{\text{txline}} = 24.969 \text{ °mil}$$

The effective dielectric coefficient is given by: (assumed const. over the design frequency band)

$$\epsilon_{\text{eff}} := \frac{\epsilon_r + 1}{2} + \frac{\frac{\epsilon_r - 1}{2}}{\sqrt{1 + 10 \cdot \frac{1}{w_{\text{'h}}(\epsilon_r, Z_0)}}} \quad \epsilon_{\text{eff}} = 3.0518$$

Therefore, the effective wavelength of a practical microstrip transmission line should be approximately:

$$\lambda_{\text{eff}}(f) := \frac{c}{f \cdot \sqrt{\epsilon_{\text{eff}}}} \quad \lambda_{\text{eff}}(836.5 \text{ MHz}) = 20.515 \text{ °cm}$$

$$\lambda_{\text{eff}}(836.5 \text{ MHz}) = 8076.9 \text{ °mil}$$

$$\lambda_{\text{e_fc}} := \lambda_{\text{eff}}(f_c) \quad \lambda_{\text{eff}}(824 \text{ MHz}) = 20.827 \text{ °cm}$$

$$\lambda_{\text{eff}}(824 \text{ MHz}) = 8199.43 \text{ °mil}$$

$$\lambda_{\text{eff}}(849 \text{ MHz}) = 20.213 \text{ °cm}$$

$$\lambda_{\text{eff}}(849 \text{ MHz}) = 7957.99 \text{ °mil}$$

The Phase Constant:

$$\beta(f) := \frac{2 \pi}{\lambda_{\text{eff}}(f)} \quad \beta(f_c) = 30.627 \text{ °} \frac{\text{rad}}{\text{m}} \quad (\text{B1.1})$$

Conductor Loss in one meter of transmission line:

$$\alpha_{\text{cond}}(f, l) := (0.072 \, \Omega) \cdot \frac{\sqrt{\frac{f}{1 \, \text{GHz}}}}{w_{\text{txline}} \cdot Z_0} \cdot l \quad \alpha_{\text{cond}}(836.5 \, \text{MHz}, 1 \, \text{m}) = 2.077 \, \text{dB}$$

Dielectric Loss in one meter of transmission line:

$$\alpha_{\epsilon}(f, l) := 27.3 \cdot \frac{\epsilon_r \cdot (\epsilon_{\text{eff}} - 1) \cdot \text{TanD}}{\epsilon_{\text{eff}} \cdot (\epsilon_r - 1)} \cdot \frac{l}{\lambda_{\text{eff}}(f)} \quad \alpha_{\epsilon}(836.5 \, \text{MHz}, 1 \, \text{m}) = 1.684 \, \text{dB}$$

Attenuation Loss Constant as a function of frequency and transmission line length:

$$\alpha(f, l) := \alpha_{\text{cond}}(f, l) + \alpha_{\epsilon}(f, l) \quad \alpha(836.5 \, \text{MHz}, 1 \, \text{m}) = 3.761 \, \text{dB} \quad (\text{B1.2})$$

$$\alpha(824 \, \text{MHz}, 1 \, \text{m}) = 3.72 \, \text{dB}$$

$$\alpha(849 \, \text{MHz}, 1 \, \text{m}) = 3.802 \, \text{dB}$$

e) Frequency / Time / Wavelength / Velocity / Delay / Phase Shift / Attenuation Relations:

$$T_c := \frac{1}{f_c} \quad : \text{Period} \quad T_c = 1.195 \, \text{ns}$$

$$\lambda(f) := \frac{c}{f} \quad : \text{Wavelength in freespace} \quad \lambda(f_c) = 35.839 \, \text{cm}$$

$$\lambda(f_c) = 14109.81 \, \text{mil}$$

Phase shift as a function of frequency f , and time delay τ :

$$\phi_{\text{shift}}(f, \tau) := \tau \cdot 2 \pi \cdot f \quad \phi_{\text{shift}}(f_c, T_c) = 360 \, \text{deg}$$

$$\phi_{\text{shift}}\left(f_c, \frac{T_c}{4}\right) = 90 \, \text{deg}$$

Phase Shift as a function of frequency f and distance d :

$$\phi_{\text{sft}}(f, d) := 2 \pi \cdot \frac{d}{\lambda_{\text{eff}}(f)} \quad \phi_{\text{sft}}(f_c, 100 \, \text{mil}) = 4.457 \, \text{deg}$$

The group velocity of the RF signal is in the transmission medium (microstrip line):

$$v_p = \lambda_{\text{eff}}(f) \cdot f \quad v_p := \frac{c}{\sqrt{\epsilon_{\text{eff}}}} \quad v_p = 6.756 \, \frac{\text{mil}}{\text{ps}}$$

$$v_p = 17.161 \, \frac{\text{cm}}{\text{ns}}$$

Distance traveled by the phase (wave front) of the signal at center frequency as a function of time delay τ at a given point:

$$d(\tau) := v_p \cdot \tau \qquad d(T_c) = 20.515 \text{ °cm}$$

The time delay in a section of RF transmission line as a function of its length l :

$$\tau_{\text{RF_Line}}(l) := \frac{l}{v_p} \qquad \tau_{\text{RF_Line}}(1 \text{ m}) = 5.827 \text{ °ns}$$

The delay in a section of RF transmission line as a function of its frequency f and phase angle ϕ :

$$\tau_{\phi}(f, \phi) := \frac{\phi}{2 \pi} \cdot \frac{1}{f} \qquad \tau_{\phi}(f_c, 90 \text{ deg}) = 298.864 \text{ °ps}$$

The transmission line length as a function of wavelength λ and phase angle ϕ :

$$\text{Length}_{\phi}(\lambda, \phi) := \frac{\phi}{2 \pi} \cdot \lambda \qquad \text{Length}_{\phi}(\lambda_{\text{eff}}(f_c), 90 \text{ deg}) = 5.129 \text{ °cm}$$

Phase shift as a function of transmission line length l :

$$\phi_{\text{TX}}(f, l) := \frac{l}{v_p} \cdot 2 \pi \cdot f \qquad \phi_{\text{TX}}(f_c, d(T_c)) = 360 \text{ °deg}$$

Phase variation with center frequency f across a bandwidth **BW** for a $\lambda/4$ transmission line:

$$\Delta\phi(f, \text{BW}) := \phi_{\text{sft}}\left(f + \frac{\text{BW}}{2}, \lambda_{\text{eff}}(f)\right) - \phi_{\text{sft}}\left(f - \frac{\text{BW}}{2}, \lambda_{\text{eff}}(f)\right)$$

$\Delta\phi(f_l, 25 \text{ MHz}) = 10.922 \text{ °deg}$	$\Delta\phi(f_l, 5 \text{ MHz}) = 2.184 \text{ °deg}$
$\Delta\phi(f_c, 25 \text{ MHz}) = 10.759 \text{ °deg}$	$\Delta\phi(f_c, 5 \text{ MHz}) = 2.152 \text{ °deg}$
$\Delta\phi(f_h, 25 \text{ MHz}) = 10.601 \text{ °deg}$	$\Delta\phi(f_h, 5 \text{ MHz}) = 2.12 \text{ °deg}$

Amplitude variation with center frequency f across a bandwidth **BW** for a $\lambda/4$ transmission line:

$$\Delta\alpha(f, \text{BW}) := \alpha\left(f + \frac{\text{BW}}{2}, \lambda_{\text{eff}}(f)\right) - \alpha\left(f - \frac{\text{BW}}{2}, \lambda_{\text{eff}}(f)\right)$$

$\Delta\alpha(f_l, 25 \text{ MHz}) = 0.017 \text{ °dB}$	$\Delta\alpha(f_l, 5 \text{ MHz}) = 3.399 \cdot 10^{-3} \text{ °dB}$
$\Delta\alpha(f_c, 25 \text{ MHz}) = 0.017 \text{ °dB}$	$\Delta\alpha(f_c, 5 \text{ MHz}) = 3.339 \cdot 10^{-3} \text{ °dB}$
$\Delta\alpha(f_h, 25 \text{ MHz}) = 0.016 \text{ °dB}$	$\Delta\alpha(f_h, 5 \text{ MHz}) = 3.28 \cdot 10^{-3} \text{ °dB}$

B2 - Main Path Insertion Loss

The main path has 4 couplers and 1 delay line, all coupled with transmission lines. When the system is off, the main path S21 results from insertion loss in components and the amount of power coupled out of the main path.

a) The main path coupler insertion loss characteristic is defined as follows:

$$S21_{CP}(CP, IL) := IL + 10 \cdot \log \left(1 - 10^{\frac{CP}{10}} \right) \quad \text{**(2.2.1.2)}$$

Yielding:

$$S21_{CP1} := -0.25 \text{ dB} + 10 \log(0.99) \quad : \text{ Insertion loss of pilot tone injection coupler (20dB)}$$

$$S21_{CP1} = -0.294 \text{ dB}$$

$$S21_{CP2} := -0.25 \text{ dB} + 10 \log(0.9) \quad : \text{ Insertion loss of noise sampling coupler (10dB)}$$

$$S21_{CP2} = -0.708 \text{ dB} \quad S21_{CP3} := S21_{CP2} \quad S21_{CP4} := S21_{CP2}$$

b) The insertion loss of the coaxial delay line is a function of its length and frequency. The attenuation characteristic can be calculated from the formula provided by the manufacturer. In the case of the main path delay line used in the FFCS:

$$S21_{coax}(f, l) := - \left[0.6283 \cdot \left(\frac{f}{1 \text{ GHz}} \right)^{0.5} + 0.04 \cdot \frac{f}{1 \text{ GHz}} \right] \cdot l \cdot \left(\frac{\text{dB}}{\text{m}} \right)$$

$$S21_{\tau_main}(f) := S21_{coax}(f, 554 \text{ mm})$$

$$S21_{\tau_main}(f_c) = -0.337 \text{ dB}$$

$$S21_{\tau_main}(f_l) = -0.334 \text{ dB}$$

$$S21_{\tau_main}(f_h) = -0.34 \text{ dB}$$

** Note: ± 0.03 dB amplitude variation across band.

c) Also considering the layout, the insertion loss of the transmission line interconnects can be calculated using (B1.2) as a function frequency f :

$$S21_{TXmain}(f) := -\alpha(f, 650 \text{ mil} + 2 \cdot 580 \text{ mil} + 2 \cdot 225 \text{ mil} + 385 \text{ mil})$$

$$S21_{TXmain}(f_c) = -0.253 \text{ dB}$$

d) Summing the insertion losses from B2 a), b) and c) yields an estimate of the total main path insertion loss as a function of frequency:

$$S21_{main}(f) := S21_{CP1} + S21_{CP2} + S21_{\tau_main}(f) + S21_{CP3} + S21_{CP4} + S21_{TXmain}(f)$$

$$S21_{main}(f_c) = -3.006 \text{ dB} \quad ; \quad S21_{main}(f_h) - S21_{main}(f_l) = 0.011 \text{ dB}$$

B3 - Amplifier Distortion Effects

a) The dynamic range according to the distortion characteristics of an amplifier may provide useful insight in the limiting factors of signal amplitude and noise on cancellation performance.

Reference: Thomas H. Lee, "The Design of CMOS Radio-Frequency Integrated Circuits," Cambridge University Press, 1998, pp. 302

b) Amplifier Characteristics (taking the FFCS Gain blocks as example)

Gain := 16.2 dB : Amplifier Gain

OIP3 := 24 dBm : Output 3rd order intercept point

P1dB := 11 dBm : Output power at 1 dB of compression

NF := 3.8 dB : Noise Figure

N_o := -174 dBm : Noise floor at room temperature

The output IP3 can be referred to the input:

$$\mathbf{IIP3} := \mathbf{OIP3} - \mathbf{Gain}$$

$$\mathbf{IIP3} = 7.8 \text{ dBm}$$

The noise floor referred to the input (input noise):

$$\mathbf{N_{oi}} := \mathbf{N_o} + \mathbf{NF}$$

$$\mathbf{N_{oi}} = -170.2 \text{ dBm}$$

The input power below IIP3 where the input-referred IM3 power equals the noise floor:

$$\mathbf{\Delta P1} := \frac{\mathbf{IIP3} - \mathbf{N_{oi}}}{3}$$

$$\mathbf{\Delta P1} = 59.333 \text{ dB}$$

The maximum input power level for spurious free operation:

$$\mathbf{P_{SFmax}} := \mathbf{IIP3} - \mathbf{\Delta P1}$$

$$\mathbf{P_{SFmax}} = -51.533 \text{ dBm}$$

The Spurious Free Dynamic Range (SFDR) of the amplifier is defined as:

$$\mathbf{SFDR} := \mathbf{P_{SFmax}} - \mathbf{N_{oi}}$$

$$\mathbf{SFDR} = 118.667 \text{ dB}$$

The above relations are expressed in **Figure B3.1**:

3rd order intermodulation (IM3) products as a function of input power (**P_{in}**):

$$\mathbf{IM3(P_{in})} := (\mathbf{N_{oi}} + \mathbf{Gain}) + (\mathbf{P_{in}} - \mathbf{P_{SFmax}}) \cdot 3$$

$$\mathbf{IM3(P_{SFmax})} = -154 \text{ dBm}$$

$$\mathbf{N_{oi}} + \mathbf{Gain} = -154 \text{ dBm}$$

$$\mathbf{IM3(0 \text{ dBm})} = 0.6 \text{ dB}$$

Signal output power of amplifier as a function of input power (**P_{in}**):

$$\mathbf{P_{out}(P_{in})} := \mathbf{P_{in}} + \mathbf{Gain}$$

The dynamic range (SNR) of the amplifier as a function of P_{in} can be expressed as:

$$SNR_o(P_{in}) := P_{out}(P_{in}) - IM3(P_{in})$$

$$SNR_o(P_{SFmax}) = 118.667 \text{ dB}$$

$$SNR_o(IIP3) = 0 \text{ dB}$$

Typical case considering the gain blocks G1 and G2:

$$P_{in1} := -33 \text{ dBm}$$

$$SNR_o(P_{in1}) = 81.6 \text{ dB}$$

$$P_{in2} := -33 \text{ dBm} + \text{Gain} - (12 \text{ dB}) \quad P_{in2} = -28.8 \text{ dBm}$$

$$SNR_o(P_{in2}) = 73.2 \text{ dB}$$

APPENDIX C - Hardware Design Notes

C1 - Cancellation Path RF Gain Blocks

The Amplifiers used are Agilents' MSA-2643. These amplifier operating conditions are calculated and scaled using the 900 MHz design specifications from the datasheet.

$f_c := 836.5 \text{ MHz}$: Design Center Frequency

The impedance of the DC block:

$$\frac{1}{2 \cdot \pi \cdot (900 \text{ MHz}) \cdot (39 \text{ pF})} = 4.534 \cdot \Omega$$

$C_{\text{Block}} := 47 \text{ pF}$ chosen for the 836.5 MHz design.

$$\frac{1}{2 \cdot \pi \cdot (f_c) \cdot (C_{\text{Block}})} = 4.048 \cdot \Omega$$

The impedance of the RF Choke:

$$2 \cdot \pi \cdot (900 \text{ MHz}) \cdot (47 \text{ nH}) = 265.779 \cdot \Omega$$

$L_{\text{Choke}} := 51 \text{ nH}$ chosen for the 836.5 MHz design.

$$2 \cdot \pi \cdot (f_c) \cdot (L_{\text{Choke}}) = 268.05 \cdot \Omega$$

The impedance of the bypass capacitor:

$$\frac{1}{2 \cdot \pi \cdot (900 \text{ MHz}) \cdot (680 \text{ pF})} = 0.26 \cdot \Omega$$

$C_{\text{Bypass}} := 740 \text{ pF}$

$$\frac{1}{2 \cdot \pi \cdot (f_c) \cdot (C_{\text{Bypass}})} = 0.257 \cdot \Omega$$

The bias stabilization resistor is calculated as follows:

$V_{cc} := 5 \text{ V}$ $V_d := 3.5 \text{ V}$ $I_d := 45 \text{ mA}$

$$R_c := \frac{V_{cc} - V_d}{I_d} \quad R_c = 33.333 \cdot \Omega$$

C2 - Charge Pump Design

The charge pump design assumes the following characteristics:

$$\beta_p := 300 \quad \beta_n := 300 \quad V_T := 26 \text{ mV} \quad : \text{ Transistor Parameters}$$

$$V_{cc} := 5 \text{ V} \quad V_{ee} := -5 \text{ V} \quad : \text{ Supply Voltages}$$

The Charge pump is controlled by the control voltage v_s that is either positive or negative

When $v_{s_p} := 3.8 \text{ V}$: Charge pump discharges the loop filter (current sink active)

$v_{s_n} := -3.8 \text{ V}$: The Charge pump charges the loop filter (current source active)

When the current source is active, the current sink is inactive and vice-versa.

The charge pump is designed to sink or source a relatively constant 5 mA current from the loop filter, to within full swing of the supply rail.

$$I_{\text{Source}} := 5 \text{ mA} \quad I_{\text{Sink}} := 5 \text{ mA}$$

To accomplished this, it is important to establish as large a voltage drop across R5,R8 as possible without saturating Q2, Q4. A voltage of approximately 300 mV is a good compromise. This allows the current source reference being generated to be relatively repeatable in the absence of good Q1, Q2/Q3, Q4 matching.

$$V_{R5} := 0.3 \text{ V} \quad V_{R8} := 0.3 \text{ V}$$

Assuming a current drive for the charge pump driver is approximately 1 mA, the voltage drop developed across R4, R9 will be approximately 258 mV considering the current density differences:

$$I_{in} := 1 \text{ mA}$$

$$V_T \cdot \ln \left(\frac{I_{\text{Source}}}{I_{in}} \right) = 41.845 \text{ mV} : \text{ Through the Q1, Q2/Q3, Q4 pairs}$$

Before finding the value of R7, it is necessary to first estimate the forward base to emitter voltage drop V_{BE_p} and V_{BE_n} of the transistors.

$$\text{Assuming: } V_{BE_p} := 0.8 \text{ V} \quad V_{BE_n} := 0.8 \text{ V}$$

Separating the pump-up and pump-down circuits facilitates nodal analysis and give the following equations:

$$R6 := \frac{(V_{cc} - v_{s_n}) - (V_{R5} + V_{BE_p})}{I_{in}} \quad R6 = 7.7 \text{ k}\Omega$$

$$R7 := \frac{(V_{ee} - v_{s_p}) - (V_{R8} + V_{BE_n})}{I_{in}} \quad R6 = 7.7 \text{ k}\Omega$$

$$R_4 := \frac{V_{R5} - V_T \cdot \ln\left(\frac{I_{Source}}{I_{in}}\right)}{I_{Source}}$$

$$R_4 = 51.631 \cdot \Omega$$

$$R_9 := \frac{V_{R8} - V_T \cdot \ln\left(\frac{I_{Sink}}{I_{in}}\right)}{I_{Sink}}$$

$$R_9 = 51.631 \cdot \Omega$$

$$R_5 := \frac{V_{R5} \cdot (\beta_p + 1)}{I_{in} \cdot (\beta_p + 1) - I_{Source}}$$

$$R_5 = 305.068 \cdot \Omega$$

$$R_8 := \frac{V_{R8} \cdot (\beta_n + 1)}{I_{in} \cdot (\beta_n + 1) - I_{Sink}}$$

$$R_8 = 305.068 \cdot \Omega$$

The final 1% resistor values used for the design were:

$$R_6 = R_7 = 7.68 \text{ k}\Omega$$

$$R_4 = R_9 = 51.1 \text{ }\Omega$$

$$R_5 = R_8 = 301 \text{ }\Omega$$

Appendix D – PCB Layout

Appendix D1 – Copper 1: RF Signal / Baseband Signal

Appendix D2 – Copper 2: RF Ground Plane, Power Ground Plane

Appendix D3 – Copper 3: DC Power +5 V

Appendix D4 – Copper 4: DC Power –5 V

Appendix D5 –VM Control Signal / Ground Plane

Appendix E – Close-up of Subcircuit Layouts and List of Components

Close-ups of sections of the PCB layout are shown with each component identified. The list of components as well as the component values are tabulated in **Table E1.1**. The resistors that appear not identified in the layout are short circuits ($\sim 0 \Omega$ resistors). Some subcircuits do not appear connected in the layout because an unused subcircuit is located in between. The subcircuits bypass the unused subcircuit with a jumping wire soldered from the output test point of one subcircuit to the input test point of the other subcircuit. For example, the output of the baseband amplifier is connected to the input of the differential to single-ended charge pump driver with a jumping wire not shown on the layout.

Appendix E1 – Copper 1: RF Signal / Baseband Signal

Figure E1.1 – FFCS RF Inputs, CP1, CP2, CP9, CP5 and G1

Shown in **Figure E1.1**: The RF input port (1) of the FFCS, the pilot tone input port (3), the pilot tone signal power divider **CP9**, the coupler **CP1** that couples the pilot tone into the main path, the noise sampling coupler **CP2** that couples the noise signal into the cancellation path, the gain block subcircuit **G1**, π -attenuator composed of resistors **R6**, **R7** and **R8**, and the vector modulator power divider **CP5**. Also shown are the RF test points **TP1**, **TP2**, **TP3** and **TP4** used to test individual subcircuits. The main path delay line port **DLP1** and pilot tone reference path delay line port **DLP3**.

Figure E1.2 – Voltage Variable RF Reflection Type Attenuators

The focus of **Figure E1.2** is on the two RF voltage variable reflection type attenuators composed of hybrid couplers **CP7** and **CP8**. The I-branch reflection type attenuator is composed of the pin diodes contained in **IC1** and the 3-dB hybrid coupler **CP6**. Also shown are RF test points **TP5**, **TP6** and **TP7** used to test individual sub-circuits.

Shown in **Figure E1.3**: I and Q-branch 3-dB hybrid power combiner **CP8**, gain block subcircuit **G2**, cancellation coupler **CP3**, error sampling coupler **CP4** and RF output port **(2)**. The I/Q demodulator subcircuit is also shown. The main path delay line port **DLP2** and pilot tone reference path delay line port **DLP4** are also shown.

Figure E1.4 – Baseband Amplifier Subcircuits

Figure E1.4 shows the baseband amplifier subcircuits. **Figure E1.5** shows the differential to single-ended amplifier that drives the charge pump circuit. The layout of the I and Q-Channel charge pumps are also shown.

Figure E1.5 – Baseband Charge Pump Drivers and Charge Pump Circuits

Figure E1.6 – PIN Diode Drivers

Figure E1.6 shows the layout of the PIN diode driver circuit.

Table E1.1 - FFCS Parts List				
Component	Company	Family	Part #	Description
RF Section				
General				
PCB	Park Nelco	N4000	N4000-13	Epoxy laminate and prepreg epoxy system (5 copper / 4 dielectric)
P1	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	RF input port (side mounted SMA connector)
P2	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	RF output port (side mounted SMA connector)
P3	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	RF pilot tone input port (side mounted SMA connector)
DLP1	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	Main path delay line input port
DLP2	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	Main path delay line output port
DLP3	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	Pilot tone reference path delay line input port
DLP4	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	Pilot tone reference path delay line output port
VCC	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	DC +5 V power port (side mounted SMA connector)
VEE	HUBER+SUHNER	SMA	82-SMA-50-0-1 / 111 NH	DC -5 V power port (side mounted SMA connector)
Main Path				
CP1	Anaren	Xinger	1D1304-20	Pilot Tone Injection 20 dB Directional Coupler
CP2	Anaren	Xinger	1D1304-10	Noise Sampling Coupler 10 dB Directional Coupler
CP3	Anaren	Xinger	1D1304-10	Cancellation Coupler 10 dB Directional Coupler
CP4	Anaren	Xinger	1D1304-10	Error Sampling Coupler 10 dB Directional Coupler
R1-R4		0603 SMD Resistor	50 Ohm	Isolated port termination
τ_{main}	HUBER+SUHNER	Sucoform	Type SM 86	Main path coaxial cable delay line (554 mm)
Gain Blocks (Cancellation Path)				
G1, G2	Agilent	MSA	MSA-2643	Low noise, cascable Bipolar MMIC Amplifier (16 dB RF Gain Stage)
C1, C2		0603 SMD Capacitor	47 pF	RF coupling capacitors
C3, C14		0603 SMD Capacitor	820 pF	RF decoupling capacitor
R5		0603 SMD Resistor	33.2 Ohm	Amplifier Bias Resistor
L1, L6		0603 SMD Inductor	56 nH	RF choke
C12, C13		0603 SMD Capacitor	47 pF	RF coupling capacitors
PI Attenuator (Cancellation Path)				
R6		0603 SMD Resistor	35.8 Ohm	
R7, R8		0603 SMD Resistor	149.9 Ohm	
Vector Modulator (Cancellation Path)				
CP5	Anaren	Xinger	4D1304	Sampled noise 3 dB in-phase power divider
CP8	Anaren	Xinger	1D1304-3	I and Q branch 3 dB hybrid power combiner
R13		0603 SMD Resistor	50 Ohm	Isolated port termination
Reflection Type Attenuators (Cancellation Path)				
IC1, IC2	Agilent	HSMP-386X	HSMP-386L	PIN diode IC
CP6, CP7	Anaren	Xinger	1D1304-3	3 dB hybrid coupler
C4, C5		0603 SMD Capacitor	47 pF	RF coupling capacitors
C6, C7		0603 SMD Capacitor	820 pF	RF decoupling capacitor
L2-L5		0603 SMD Inductor	56 nH	RF choke
R9-R12		0603 SMD Resistor	43.2 Ohm	PIN diode control bias network
C8, C9		0603 SMD Capacitor	47 pF	RF coupling capacitors
C10, C11		0603 SMD Capacitor	820 pF	RF decoupling capacitor
I/Q Demodulator				
CP9	Anaren	Xinger	4D1304	Pilot Tone signal 3 dB in-phase power divider
IC3	MACOM	E-Series SMT	ETC1-1-13	Single-to-differential transmission line transformer
IC4	Analog Devices		AD8347	0.8 - 2.7 GHz Direct Conversion Quad Demod
C15, C16		0603 SMD Capacitor	100 pF	LO port coupling capacitors
R15, R16		0603 SMD Resistor	200 Ohm	LO and RF port matching resistor
C19, C20		0603 SMD Capacitor	100 pF	RF port coupling capacitors
C17, C21, C26		0603 SMD Capacitor	100 pF	Decoupling capacitors
C18, C22, C23, C24, C25, C27		0603 SMD Capacitor	0.1 μ F	Decoupling capacitors
τ_{ref}	HUBER+SUHNER	Sucoform	Type SM 86	Pilot tone reference path coaxial cable delay line (1007 mm)

Table E1.1 - FFCS Parts List (continued...)				
Component	Company	Family	Part #	Description
Baseband Section				
General				
C29, C31, C33, C35, C36, C38, C40, C42, C50, C52, C56, C59		T492-C Capacitor	10 μ F	Tantalum decoupling capacitor
C28, C30, C32, C34, C37, C39, C41, C43, C51, C53, C57, C58		0603 SMD Capacitor	0.1 μ F	Decoupling capacitors
Baseband Amplifiers				
IC5, IC6	Analog Devices		AD8132	High-speed Differential Amplifier
R17, R18, R21, R22		0603 SMD Resistor	475 Ohm	
R19, R20, R23, R24		0603 SMD Resistor	4.75 kOhm	
Charge Pump Driver				
IC7, IC8	Analog Devices		AD9631	Ultra low distortion, wide bandwidth voltage feedback opamp
R25, R26, R29, R30		0603 SMD Resistor	51.1 Ohm	
R27, R28, R31, R32		0603 SMD Resistor	100 kOhm	
Charge Pump				
IC9, IC11	Philips	BC8xx	BC857BS	General purpose PNP transistor pair
IC10, IC12	Philips	BC8xx	BC847BS	General purpose NPN transistor pair
R33, R34, R40, R41		0603 SMD Resistor	7.68 kOhm	
R35, R36, R42, R43		0603 SMD Resistor	51.1 Ohm	
R37, R38, R44, R45		0603 SMD Resistor	301 Ohm	
C45, C47		0603 SMD Capacitor	2 μ F	Loop Filter / Charge Storage capacitor
PIN Diode Driver				
IC13, IC14	Analog Devices		AD9631	Ultra low distortion, wide bandwidth voltage feedback opamp
R47, R51		0603 SMD Resistor	619 kOhm	
R50, R54		0603 SMD Resistor	750 kOhm	
R49, R53		0603 SMD Resistor	100 Ohm	