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Components and Architectures for an Integrated Wideband FM Transceiver

by

Terrence P. Kenny, B.Sc. (Eng.)

A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of
the requirements for the degree of
Master of Engineering

**Ottawa-Carleton Institute for Electrical Engineering
Department of Electronics
Faculty of Engineering
Carleton University
Ottawa, Canada
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**COMPONENTS AND ARCHITECTURES FOR AN
INTEGRATED WIDEBAND FM TRANSCEIVER**

submitted by
Terrence Patrick Kenny, B.Eng.

in partial fulfilment of the requirements
for the degree of Master of Engineering.

Chair, Department of Electronics

Thesis Supervisor

Carleton University

January 1994

The information used in this thesis comes in part from the research program of Dr. Miles A. Copeland and his associates in the VLSI in Communications group. The research results appearing in this thesis represent an integral part of the ongoing research program. All research results in this thesis including tables, graphs and figures but excluding the narrative portions of the thesis are effectively incorporated into the research program and can be used by Dr. Copeland and his associates for education and research purposes, including publication in the open literature with appropriate credits. Matters of intellectual property may be pursued collaboratively with Carleton University and Dr. Copeland where and as appropriate.

Abstract

This thesis reports on the work performed in the experimental validation of a digitally controlled broadband coded FM concept, based upon the use of a $\Delta\Sigma$ frequency synthesizer as both a transmitter and a programmable local oscillator, and a $\Delta\Sigma$ frequency discriminator as a part of an FM demodulator. The concept and components investigated would support a high degree of VLSI integration in the transceiver. The design and test results of alternative printed circuit board prototypes are presented and a description is given of how these components were used in the investigation of a broadband digital FM transceiver. Experimental validation of the concept was addressed with the use of a single Fourier component, a digitally encoded sinusoidal FM sequence, varied to produce changes in both frequency deviation and repetition rate. Sideband noise at IF frequency, a product of the conversion of two broadband driven synthesizers, was measured as a function of the repetition rate and peak deviation. The suitability of the output result for $\Delta\Sigma$ frequency discrimination was evaluated. Finally, the proposed transceiver concept was reviewed in light of these experiments.

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List of Symbols and Abbreviations

A/D	analog-to-digital
β_{FM}	peak phase deviation
CLB	configurable logic block
DFT	discrete Fourier transform
DMD	dual modulus divider
DPA	digital phase accumulator
DSP	digital signal processing
Δf	sinusoidal FM deviation
$\Delta\Sigma$	delta-sigma
$\Delta\theta_{e,m}$	synthesizer PLL phase error mismatch
ECL	emitter coupled logic
EMI	electro-magnetic interference
<i>ENBW</i>	equivalent noise bandwidth
EPROM	eraceable and programmable read only memories
F	linear representation of noise figure
FFT	fast Fourier transform
f_m	sinusoidal FM rate
FM	frequency modulation
FPGA	field programmable gate array
G	power gain

IF	intermediate frequency
IM	image frequency
K_D	discriminator constant
LCA	logic cell array
LNA	low noise amplifier
LO	local oscillator
LPF	low pass filter
LSB	least significant bit
MASH3	a third order $\Delta\Sigma$ modulator composed of three first order cascaded structures
MEC	modulus extension circuit
NF	noise figure
OIP_3	third order intercept point
P_{1dB}	1 dB compression point
PAL	programmable array logic
PBX	private branch exchange
PCB	printed circuit board
PFD	phase-frequency detector
PLL	phase locked loop
PN	pseudo-random noise
RAM	random access memory
RF	radio frequency
ROM	read only memory
R_x	receiver
SAW	surface acoustic wave
SNR	signal-to-noise ratio

SQNR	signal-to-quantization-noise ratio
T_x	transmitter
$\theta_b(t)$	phase variation on the DMD introduced by the test vector
UGBW	unity gain bandwidth
VCO	voltage controlled oscillator
VLSI	very large scale integration
ω_n	natural loop frequency
ζ	damping coefficient

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Chapter 1

Introduction

The realization of a digital personal communications radio transceiver operating in the GHz range that meets the specifications of an industry standard is a challenging effort that requires understanding of both component and system level design. Partial motivation for this work is the investigation of alternative component architectures that allow a higher level of VLSI integration, including digital frequency control and a greater level of integrated digital processing. Additional motivation exists for investigation of a broadband pseudorandom coded FM application that may be suitable in combatting the effects of multipath fading.

Possibilities for integrated transceiver architectures are explored in this thesis through the construction of two discrete prototype components that can serve as building blocks in higher level radio design. A delta-sigma ($\Delta\Sigma$) based fractional-N PLL frequency synthesizer [RCK93] is one component that is investigated. Its usefulness as a digitally programmable local oscillator makes it indispensable in the design of multi-channel angle modulated receivers. Equally vital is its usefulness as a digitally controlled angle modulating transmitter. The digital delta-sigma modulator driving the synthesizer provides frequency programmability with a resolution dependant upon the number of bits on its input and the reference frequency used. A second component, a digital delta-sigma frequency discriminator [BC92], is also investigated. This component is capable of converting analog frequency deviation into an oversampled digital single bit bitstream whose density of ones is proportional to the frequency deviation of the input signal. This makes the component suitable for a receiver that requires A/D conversion and demodulation of a wide deviation low data rate FM modulated signal.

This thesis studies these two components and proposes an alternative digital radio receiver, as part of a transceiver architecture, which incorporates their features. A

test bed is developed, which makes use of a broadband sinusoidal FM test vector, for verification of the proposed architecture. The broadband concept is motivated by the need to address the frequency selective fading impairment inherent in mobile radio transceivers. As well, the broadband application appears well suited to the two radio components investigated in this thesis.

1.1 Thesis Outline

This thesis is organized over six chapters. Chapter 2 provides an introduction to the architecture of an FM receiver. The capability of digital channel programmability is addressed by the introduction of a $\Delta\Sigma$ frequency synthesizer as a controllable LO. Analog to digital conversion and digital data recovery, at the IF stage, are considered with the use of a $\Delta\Sigma$ frequency discriminator. The configuration is suitable for both narrowband FM and broadband coded FM. An architecture for a broadband digitally encoded FM receiver is proposed and an experimental testbed described for evaluation of the broadband coded FM concept. Chapter 3 presents the design and analysis of a digital delta-sigma frequency synthesizer. Quality issues such as quantization noise, phase noise and transient response are addressed through the use of simulation and measurement. The design and analysis of a second order $\Delta\Sigma$ frequency discriminator is presented in chapter 4. Measurements of SNR, effective bit resolution and quantization noise shaping indicate the quality of operation. Chapter 5 provides a set of measurements that characterize the operation of the broadband coded FM application presented in chapter 2 and using the component previously discussed. Chapter 6 presents a conclusion for the work done, and makes some recommendations for continuing work on this type of broadband receiver.

Chapter 2

A Radio Receiver Architecture

In this chapter, the general background of an FM radio receiver design is reviewed. The particular context of an FM radio receiver is then introduced using the alternative components of this thesis.

Radio receiver designs are dependant upon channel impairments and the parameters specified for a particular application. In this chapter, after a review of some basic concepts, a digital FM receiver architecture is presented based upon the use of a digital $\Delta\Sigma$ frequency synthesizer [RCK93] as a controllable LO, and a digital $\Delta\Sigma$ frequency discriminator [BC92] as part of the IF demodulator. The architecture appears well suited for broadband coded applications which continuously move (slither) a carrier signal in a pseudo-random manner over a given spreading bandwidth. The broadband carrier in conjunction with some form of diversity combining in the receiver will provide benefits in overcoming the effects of frequency selective fading, the primary impairment of a wireless environment.

2.1 An FM Radio Receiver Design

In FM systems, generation of the transmit signal is usually accomplished with a voltage to frequency converting device (ie. phase locked VCO), where a voltage varying input $m(t)$ results in a frequency modulated output $v(t)$.

$$v(t) = A \cos \left[\omega_c t + k \int_{-\infty}^t m(t) dt \right] \quad (2.1)$$

For a sinusoidally varying modulation, signal $v(t)$ can be redefined as

$$v(t) = A \cos [\omega_c t + \beta_{FM} \sin \omega_m t] \quad (2.2)$$

Here β_{FM} , the maximum phase deviation, is referred to as the modulation index, and is a function of the maximum frequency deviation Δf and modulation rate f_m , and is given by

$$\beta_{FM} = \frac{\Delta f}{f_m} \quad (2.3)$$

The instantaneous frequency is defined as

$$f = f_c + \beta_{FM} f_m \cos \omega_m t \quad (2.4)$$

This indicates that the instantaneous frequency can always be found within $f_c \pm \Delta f$, yet the Fourier expansion of the modulating term of equation (2.2) results in a spectrum consisting of a carrier and a set of sidebands symmetrically spaced on each side of the carrier (ie. Bessel functions). Theoretically, for an FM modulated signal the number of sidebands is infinite, requiring an infinite bandwidth. For all practical purposes for all values of β_{FM} this bandwidth is substantially reduced because a greater portion of the total power is contained in a limited number of sidebands. Experimentally it has been shown [TS86] that distortion resulting from bandlimiting an FM signal can be acceptable if at least 98% of the total power is transmitted through the bandlimiting filter. Therefore, for a sinusoidally modulated FM signal, the required bandwidth B is described by

$$B = 2(\beta_{FM} + 1)f_m = 2(\Delta f + f_m) \quad (2.5)$$

which is the well known Carson's rule.

For wideband FM, generally the Fourier expansion of the frequency modulated waveform cannot be evaluated. The several mechanisms which define the spectrum of an FM waveform [Str79] are the rate of change of the modulating signal and the ratio between amplitude of the modulating signal and the

instantaneous frequency of the FM signal. Although the instantaneous frequency follows the amplitude of the modulating signal, the spectral density may not result in the same pattern. This occurs because the definition of instantaneous frequency and the frequency used in the Fourier transform are not the same. The value of β_{FM} plays a large role in determining the bandwidth requirements for an FM system. If β_{FM} is small, then for sinusoidal FM, the spectral content of the modulating signal

Figure 2.1 : Approximation to the magnitude FM spectral density as β_{FM} goes to infinity, for the sinusoidal case. (reprinted from [Str79])

predominantly controls the magnitude of the FM spectral density requiring a bandwidth of $2f_m$. If β_{FM} is quite large then amplitude to frequency conversion predominantly controls the amplitude of the spectral density leading to a bandwidth requirement of $2\Delta f$. Figure 2.1 shows the frequency output signal with respect to time and the expected output spectrum for a wideband sinusoidally modulated FM signal.

Figure 2.2 shows the block diagram of a generalized single conversion FM receiver. These type of receivers are designed to provide maximum signal to noise ratio from an RF input that varies widely in signal strength, and may contain adjacent channel interference.

Upon reception at the antenna the RF signal is processed by a low noise amplifier (LNA). The LNA is used as an initial amplifying stage to boost the power of the received signal while minimizing the addition of noise and distortion. Large gain in the LNA leads to greater overall signal to noise ratio (SNR) if subsequent amplification stages are used in the system. Desirable characteristics for the LNA

are large power gain, low noise figure, a wide dynamic range linear transfer function, dynamic stability, reverse isolation, and frequency selectivity (see Figure 2.7). The operating characteristics for the LNA are usually a compromise between these options [KBR80]. The LNA can be preceded by a wideband matching network that reactively matches the source admittance for optimum noise performance. The overall noise figure F (the linear representation of NF) of the LNA then becomes a function of both the active devices used and the source admittance, and is given by [NS93]

$$F = F_{min} + \frac{R_n}{G_g} \left[(G_{on} - G_g)^2 + (B_{on} - B_g)^2 \right] \quad (2.6)$$

Figure 2.2 : Block diagram of a typical FM receiver.

Here $(G_g + jB_g)$ is the generator admittance presented to the input of the LNA, and $(G_{on} + jB_{on})$ the generator admittance that provides optimum noise performance. R_n is the noise resistance and indicates the sensitivity of F to generator admittance, while F_{min} is an intrinsic property of active devices.

For bipolar type amplifiers the factor which determines noise levels is the input impedance match. For each current/frequency operating point of the amplifier there is an optimum input match $(\Gamma_{op} = R_{op} + jX_{op})$ for minimum F . If F is presented as a function of reflection coefficients then

$$F = F_{min} + \frac{4R_n}{Z_o} \frac{|\Gamma_s - \Gamma_{op}|}{|1 + \Gamma_{op}|^2 \cdot (1 - |\Gamma_s|^2)} \quad (2.7)$$

and for an optimum input match ($\Gamma_s = \Gamma_{op}$) minimum F occurs. Here Γ_s is the reflection coefficient seen by the input of the LNA.

For a silicon bipolar single transistor or cascode type amplifier the output of the LNA can be matched for optimum power gain using a tank circuit on the collector of the output transistor. This matching network can also incorporate an image reject filter to suppress the noise at the image frequency, thereby ensuring better SNR at the IF stage.

Frequency translation can be accomplished using active or passive mixers. For discrete versions the passive diode mixer is the preferred choice. However, most silicon RF VLSI implementations use the bipolar Gilbert cell because of the inherent conversion gain, isolation and harmonic suppression and because it can be integrated more easily. When preceded by a LNA, under proper matching conditions, the frequency translation can be achieved with greater dynamic range. The RF input on the bipolar mixer sees the output impedance of the LNA in combination with the image reject filter. For a cascode amplifier the output impedance is relatively high, and therefore wideband impedance matching between the LNA and mixer can be achieved by placing a 50Ω shunt resistor across the high impedance mixer input to Vcc or Gnd. This helps to widen the 3 dB bandwidth of any parasitic low pass filter present on the mixer input by moving the pole further from DC. This same technique can also be used on the LO input port. If an emitter follower buffer is used at the mixer output then a 25Ω resistor can be placed in series with its low output impedance to provide some degree of wideband output impedance matching under varying load conditions.

The LO input on the mixer is sourced by an oscillator that is matched for maximum power transfer to ensure minimum reflection, corruption, and harmonic distortion of the oscillator signal. The most important figure of merit for the LO is double sideband noise. This phase noise on the LO will be translated directly to the IF output at the same ratio as it appeared on the LO. For multichannel receivers this can present a SNR problem if an adjacent channel, transmitting nearby, has large levels of sideband noise at a channel spacing offset. For a linear oscillator the content of this sideband noise for RF or microwave oscillators is to a great extent

phase noise, but may also contain $1/f$ (flicker) noise, externally coupled noise and microphonic (mechanical vibration) noise. The level of phase noise in an oscillator, per unit frequency offset, is a component value that is dependant upon many factors. The phase noise to carrier power ratio for a linear oscillator using a non-ideal amplifier is given by [Rob91]

$$\left(\frac{N}{P}\right)_{f_{os}} = \frac{FkT}{P} \cdot \frac{1}{8Q^2} \cdot \left(\frac{f_c}{f_{os}}\right)^2 \quad (2.8)$$

Here F (the linear representation of NF) is a function of the output power level P . Therefore the phase noise levels are a function of the noise figure, the output power level, the loaded Q of the feedback elements in the oscillator circuit and the corner frequency f_c of the active device.

The output of the mixer is followed by an IF filter which is used to further suppress adjacent channel interferers or other unwanted signals. The demodulation stage should be presented a signal with the specified in-band SNR and out-of-band suppression. Filters used at this stage are typically high Q components used to pass single channel information at the given IF frequency and modulation bandwidth. They may come in the form of SAW, crystal, ceramic, or active tuned filters.

Demodulation of an FM signal can be accomplished with a magnitude limiting frequency to voltage converter where the output voltage is proportional to the frequency deviation of the input signal. However a digital output is desirable to enable greater levels of integrated digital signal processing (DSP).

2.2 A $\Delta\Sigma$ Digital FM Radio Receiver

A digital $\Delta\Sigma$ FM radio receiver is defined, in this thesis, based upon the use of a $\Delta\Sigma$ fractional-N PLL frequency synthesizer as a programmable local oscillator, and a $\Delta\Sigma$ frequency discriminator to provide A/D conversion and frequency discrimination for the FM demodulation process. A $\Delta\Sigma$ fractional-N PLL frequency synthesizer is a traditional PLL synthesizer that replaces the fixed divider in the feedback path with a dual modulus divider. Using a $\Delta\Sigma$ modulated bitstream to control the division modulus allows the synthesis of fractional values of frequencies. The $\Delta\Sigma$ frequency discriminator is a novel architecture by Beards

and Copeland [BC92] that converts an analog frequency modulated signal to a coarsely quantized digital signal with the modulation signal embedded as a bit density in the bitstream output.

The digital control on both the $\Delta\Sigma$ frequency synthesizer and the $\Delta\Sigma$ frequency discriminator opens up opportunities for exploring precisely controlled FM architectures for analog/digital radio. By their very nature each component is suited to wide frequency deviation (>100 KHz), low data rate (<100 KHz) FM applications. For this reason the thesis work focuses the investigation on their suitability for broadband FM.

Figure 2.3 shows the block diagram of an FM radio receiver implementation using a $\Delta\Sigma$ frequency synthesizer for the LO, and a $\Delta\Sigma$ frequency discriminator for part of the FM demodulation. For narrowband FM the synthesizer can provide the required frequency programmability for a multi-channel receiver, although the issue of narrowband FM will not be addressed in this thesis. The precise frequency control of the synthesizer is dictated by the number of frequency control bits on the $\Delta\Sigma$ modulator input as well as the value of its reference frequency. Alternatively, if the control bits were changed at a rate slower than the transient response of the PLL, then the receiver could be used as a central component for frequency hopping or broadband coded FM applications, i.e.; it could be doing the broadband down conversion. For an n bit interpolated pseudo-random noise (PN) code on the synthesizer $\Delta\Sigma$ modulator input, the useful frequency deviation for broadband movement would be dictated by the useful average bitstream density β (fraction of ones per unit section of serial bits) of the output of the $\Delta\Sigma$ modulator. Third order modulators are characterized by useful values of β between 0.25 and 0.75, and therefore the useable programmable bandwidth becomes

$$\omega_{ss} = 2\pi(\beta_{max} - \beta_{min})f_{ref} \quad (2.9)$$

where β_{max} and β_{min} are the upper and lower limits of average bitstream density, respectively.

It is interesting to note that the frequency deviation associated with broadband frequency control could be considerably higher than the modulating frequency rate, which is limited by the PLL loop bandwidth.

Figure 2.3 : Block diagram of a $\Delta\Sigma$ Digital FM receiver.

The $\Delta\Sigma$ frequency discriminator converts the analog IF information to a digital bitstream that has high-pass quantization noise shaping. This output signal, an oversampled bitstream, contains the embedded modulation data. The rate of this oversampled output is equal to the reference frequency clock driving both the discriminator and the $\Delta\Sigma$ modulator. For high reference clock speeds, and hence oversampling rates, wideband frequency demodulation can be accomplished. Post discrimination decimation filtering can be used to remove the quantization noise as well as reduce the oversampled rate to the baseband data rate. The problems of synchronizing the R_x frequency synthesizer to the spreading code of the T_x signal would be solved in the processing of the digital output of the frequency discriminator and are not dealt with in this thesis. A further thesis project is pursuing this issue [Fi94].

2.2.1 A General Broadband FM T_x/R_x Test Set-up.

Figure 2.4 (a) shows the general block diagram of a broadband FM transmit/receive channel which the work in this thesis is aimed at. At the transmitter, data signal $f_{s1}(t)$ is added through mixing to broadband spreading signal $f_{bb1}(t)$, producing transmit signal $f_{tx}(t)$. If an ideal channel characteristic is assumed, then at the receiver an estimation of the transmitted broadband code $f_{bb2}(t)$ can be subtracted in a frequency sense from the received signal (i.e. mixed), producing

signal $f_{IF}(t)$. If both the transmit and receive broadband spreading signal frequencies are well matched and the addition and subtraction operations are considered linear, then the signal recovered at the IF will be a good approximation of the transmitted signal $f_{si}(t)$.

$$f_{IF}(t) = f_{si}(t) - [f_{bb2}(t) - \hat{f}_{bb1}(t)] \cong f_{si}(t) \quad (2.10)$$

(a)

(b)

(c)

Figure 2.4 : a) A general T_x/R_x setup for broadband FM;
 b) representation of a broadband sinusoidal FM transmit signal;
 c) representation of the approximated received signal.

If the IF stage is followed by the $\Delta\Sigma$ frequency discriminator, then the discriminator output will be an oversampled single bit bitstream with a bit density proportional to the IF frequency. The discriminator, by the nature of its design, is suited to wide frequency deviation FM signals. Therefore a large frequency deviation testbed, compatible in frequency range with the discriminator, would allow synchronized testing of the broadband FM concept.

The fundamental issue requiring investigation is the quality of IF signal resulting from the broadband mixing of the transmit and receive frequencies. For the proposed architecture to be practical it is essential that the quality of this IF signal be compatible with the signal requirements of the $\Delta\Sigma$ frequency discriminator, which favors wide frequency deviation signals (directly related to SNR).

2.2.2 A Broadband Coded FM Application

Figure 2.5 shows the block diagram of a receiver architecture that could be used for broadband coded (frequency slithering) FM applications. The transmit carrier frequency is caused to move over a wide frequency deviation range, sinusoidally or pseudorandomly, except the movement is a continuous interpolation (filtered response) and therefore does not stress the transient behavior of the PLL of the frequency synthesizer in the receiver. This provides the receiver with the opportunity to follow the transmit frequency with a desired fixed frequency offset (to mix to the IF frequency) and, under ideal conditions, with a minimum level of IF sideband noise. The general T_x/R_x setup shown in Figure 2.4 allows exploration of the feasibility of this architecture. For non-strategic broadband frequency environments the complexity of the PN code and the receiver circuit is reduced because detection by unauthorized users, or SNR degradation by intentional interferers, is not an issue. This obviates the use of some form of transmitter preamble to designate channel position and therefore alleviates the burden of coarse acquisition. If a microcellular environment is considered, such as one used for an indoor wireless PBX, then each user could transmit on the same filtered spreading PN code but be assigned a specific phase or bit position within the code. The PBX base station would continuously transmit on all code positions. When a channel or code position was unused, the base station would transmit information about being ready for access as well as the start address for the channel code phase position. In this way each portable receiver, through the use of the coarse synchronization circuit, would then be supplied with a code phase position which would be used by the address generator to pre-align the PN code to the proper sequence. This sequence, in conjunction with the fine synchronization control circuit, would provide an accurate estimate of the transmitted code phase position. The maximum number of users that could be accommodated would be determined by considering the length of the PN code, the number of bits required per user, and the number of bits required to separate each user to ensure minimum cross

Figure 2.5 : Block diagram of a broadband coded FM receiver.

modulation. Doppler shifts introduced by indoor radio environments tend to be small with respect to digital FM offsets and therefore the frequency bins could be designed to accommodate these shifts. Therefore choice of the spreading PN code length and its rate (with respect to the data modulation rate and frequency deviation) would aid in the design of a receiver with a maximized SNR and a minimized correlation with the nulls of the slow frequency selective fading resulting from the delay spread of the indoor channel. The advantages of a PN code over a simple sinusoidal FM carrier is the resultant uncorrelated sideband noise occurring at the IF. With each user switching on a different position of the same code, care would have to be given to code design to minimize frequency crossover (collisions) of multi-access users. Under normal operating conditions most interference would be the result of channels coming on line, or adjacent cell users.

The use of feedback to the transmitter to control FM deviation of the transmitted data could be used to increase SNR and therefore minimize these effects.

An alternative architecture from the above, using a LNA and $\Delta\Sigma$ frequency discriminator (no broadband conversion mixer), would be possible for cordless telephone applications where a single user per microcell is assumed. The signal correlation would be done digitally after the A/D conversion, in the discriminator. The main disadvantage of this architecture is the inability of the discriminator to resolve low level frequency signals in the presence of large magnitude interferers. This drawback limits the general purpose use of this architecture for digital radio.

2.2.3 A Test Bed for Experimental Validation of Broadband Coded FM.

In this thesis the design and implementation issues of (coarse and fine) synchronization are not dealt with. Fine synchronization in the testbed is achieved through the use of a common reference signal supplied to both synthesizers and the data generator. Figure 2.6 shows a general block diagram of the testbed structure. The upper left synthesizer in the Figure models part of the transmitter, the bottom right synthesizer part of the receiver. The testbed therefore realizes the general T_x/R_x set-up discussed in section 2.2.1 (Figure 2.4), with $f_{bb1}(t)$ and $f_{bb2}(t)$ being identical except for analog discrepancies between the two synthesizers. The T_x synthesizer uses a direct digital control technique (digital $\Delta\Sigma$ modulator inputs) for producing the combined broadband frequency and FM modulation frequency ($f_{bb1}(t) + f_{\alpha}(t)$ from Figure 2.4). In the test bed diagram of Figure 2.6 the T_x mixer and offset frequency are present only for the purpose of generating a desirable RF transmit frequency (a pragmatic solution for testing), because typically the T_x synthesizer would generate broadband frequencies, f_{TX} , that are always an IF frequency away from the broadband downconversion frequencies at the receiver, while in the testbed setup identical synthesizers are used which cannot be so separated. The R_x synthesizer generates only a despreading code for the purposes of data recovery. Combined, the T_x and R_x functions indicate the nature of the proposed channel, although the T_x synthesizer is not the only possible way of generating the T_x signal. As well, the combination here indicates how a transceiver might be built, ie; how the transmit function in a transceiver might be achieved.

This could imply, for instance, full duplex communication in, for example, wireless modem.

The analog differences between the T_x and R_x synthesizers will lead to 'frequency impairments' at the IF stage which are a function of the bandwidth limitations of the synthesizers, the disparities between the T_x/R_x VCOs, phase noise levels, and the frequency response of the R_x mixer. The test set-up models the consequences of such effects. It may not duplicate exactly how these impairments arrive in the T_x/R_x combination; for example the final transceiver combination would not require a T_x mixer and offset frequency. However by proper programming of the T_x/R_x synthesizers in the test set-up it will be possible to demonstrate in a general way the consequences of the impairments mentioned above. This will help to establish if this is a viable receiver concept. Thus the issue being investigated, through the use of the testbed, is the nature of the signal resulting from the conversion of the exact frequency differences of a combined $[f_{si}(t)$ and $f_{bb1}(t)]$ with $f_{bb2}(t)$ as defined in Figure 2.4.

One technique for measuring minor frequency variations between the T_x/R_x synthesizers would be an evaluation of the signal strength of the IF tone (chosen to be 105 MHz) with respect to sideband noise and distortion. In this case the 105 MHz IF frequency was chosen arbitrarily, although higher frequencies are possible. Thus, assuming the acquisition problem is solved, defining the level of the sideband noise is in effect characterizing the potential quality of the proposed receiver.

Because identical T_x and R_x synthesizers are used, along with an IF frequency on both of the T_x and R_x sides, a single test vector can be used to drive both the T_x and R_x synthesizers. The test vector does not need to contain FM data or need to be moved pseudorandomly in frequency to perform the characterization, but needs to cover the extended frequency range. The experimental test vector chosen to drive both the transmit and receive frequency synthesizers is a single time varying component of a general Fourier series, encoded as a double precision digital sinusoid. This provided a pragmatic technique, similar to the one used by Sorfleet [Sor91], for testing the broadband concept. A bitstream consisting of double precision $\Delta\Sigma$ noise shaping code, modulated by a wideband FM sinusoidal vector was produced in software [Ri2] and stored as an ASCII file. Using a bitstream generator card (designed for the XT style bus of an IBM compatible computer), the stored bitstream was loaded from a file on floppy disk to RAM on the

Figure 2.6 : Block diagram of the experimental testbed for broadband coded FM.

generator card and then shifted out serially from the card to the DMD (and MEC) of the frequency synthesizer, synchronized to a master 10 MHz clock. Figure 2.6 shows the block diagram of the testbed. One limitation of this technique is that it is characterized by limit cycles because of the one million bit finite sequence code length. For a 10 MHz clock rate and a 10^6 bit sequence length, a 10 Hz tone would develop. However, a low level tone of this frequency would not seriously impair testing.

Variation of two independent parameters, the FM repetition rate and FM deviation, allowed the desired characterization of the broadband coded FM concept.

2.2.4 Motivation for the Experimental Test Bed Model

In the experimental set-up of Figure 2.6, in order to validate that broadband coded FM is a viable concept, the level of SNR with respect to sideband tones in the final receiver IF signal must be quantified. For the purpose of initially qualifying the broadband FM concept it is not necessary to introduce a set of random data into the experiment.

Sideband tones in the test set-up and the proposed transceiver concept can arise for several reasons. These include synchronization imperfections between the clocks driving the Tx and Rx $\Delta\Sigma$ modulators, and impairments in the path between the Tx and Rx frequency synthesizer outputs. In the test set-up, ideal synchronization has been assumed (by the use of a common reference frequency). In a true receiver some form of fine synchronization is always implied as shown in Figure 2.5; the equivalent of this is provided in the test bed set-up by direct connection. If ideal fine time synchronization for the receiver is assumed (which will be shown to be reasonable) then the path impairments between the Tx synthesizer output and the receiver IF signal must be qualified, because these impairments will directly affect the sensitivity (minimum detectable sidetone level) of the broadband measurement.

2.3 Figures of Merit for Receiver Performance

In this section a review of some front end receiver issues is included for completeness, although the front end requirements are not considered further in this thesis.

Some of the fundamental criteria for evaluating receiver performance include power gain G , noise figure NF , third order intercept point OIP_3 , 1 dB compression point P_{1dB} , image frequency IM , receiver phase noise, and linear (spurious free) dynamic range [Wat86,KBR80]. Other measures of performance can usually be derived from these.

G can be defined as the ratio of the power delivered to the load by the available source power, and is given by

$$G = \frac{P_{out}}{P_{in}} = \frac{V_{out}^2/R_L}{V_{in}^2/4R_S} = 4 \frac{R_S \cdot V_{out}^2}{R_L \cdot V_{in}^2} \quad (2.11)$$

where P_{in} is defined as the power across the input impedance Z_{in} of the amplifier. When Z_{in} is not equal to R_S then maximum power is not transmitted, hence maximum gain cannot be achieved.

NF , a measure of noisiness, can be defined as the degradation of signal-to-noise ratio between the input and output for a two port device [ZT85].

$$F = \frac{SNR_i}{SNR_o} = \frac{S_i/N_i}{(G \cdot S_i)/(N_a + G \cdot N_i)} = \frac{N_a + G \cdot N_i}{G \cdot N_i} \quad (2.12)$$

$$NF_{dB} = 10 \log (F) \quad (2.13)$$

This can be an indication of additional increases in the noise floor of the system if electronic or thermal noise is considered. In a receiver context, NF then becomes a measure of the degradation of the sensitivity or dynamic range of the system. If a cascade connection of gain stages are involved, such as in a receiver with an active mixer preceded by a LNA and followed by an IF gain stage, then the NF of the latter stages are suppressed by the gain of the preceding stages.

$$F_{tot} = F_{LNA} + \frac{F_{Mixer} - 1}{G_{LNA}} + \frac{F_{IF} - 1}{G_{LNA} G_{Mixer}} + \dots \quad (2.14)$$

NF measurement can be accomplished by varying the power level of a noise source connected to the input of the device under test, and then measuring the level of the output noise power using a receiver.

Harmonic and intermodulation distortion are indications of the level of non-linearity of a component or device. If a power series description for the transfer function of a non-linear device is considered [PM91] then the output signal is given by

$$v_o = a_1 v_i + a_2 v_i^2 + a_3 v_i^3 + \dots \quad (2.15)$$

Harmonic distortion results from the passing of a time varying signal (v_i) through a non-linear device. The output spectrum will contain a scaled version of the fundamental signal plus harmonics or second order terms. Intermodulation distortion results from the passing of two or more time varying signals through a non-linear device. The output spectrum will contain both fundamentals and their

Figure 2.7 : Input power vs. output power response for a non-linear device.

harmonics as well as the beat frequencies of the fundamentals and their harmonics. If the input signals are constrained to be sinusoids and v_i is defined as

$$v_i = V_1 \cos \omega_1 t + V_2 \cos \omega_2 t \quad (2.16)$$

then the output spectral terms are dc , f_1 , f_2 , $f_1 \pm f_2$, $2f_1$, $2f_2$, $2f_1 \pm f_2$, $2f_2 \pm f_1$, $3f_1$, $3f_2$, ... etc. The terms $2f_1 \pm f_2$ and $2f_2 \pm f_1$ are the third order intermodulation products, and their difference terms are very important because, being close to the fundamentals, they can fall within the bandwidth of concern and lead to distortion in the output signal. For this reason the resulting power levels of the third order intermodulation products as a function of the fundamental power levels are very useful measurements for characterizing the quality of receiver components as well as determining appropriate operating power levels. The OIP_3 , as shown in Figure 2.7, is defined as the point where the third order intermodulation product equals the uncompressed linear output, requires several *RF* inputs for characterization. The P_{1dB} , defined as the point where the measured gain is 1 dB below the ideal linear gain, requires only a single *RF* input. Because the P_{1dB} and OIP_3 are mathematically related, the easier P_{1dB} measurement is usually sufficient in

qualifying linearity. The simplified relationship between OIP_3 and P_{1dB} is given by

$$OIP_3 \cong P_{1dB} + 10dB \quad (2.17)$$

The image frequency IM is defined as a signal found on the opposing side of the LO from the RF input, and offset from the LO by exactly the IF frequency. Through the act of linear mixing both the RF and IM signals can be translated to the IF band. Signal degradation arising from image frequencies is usually removed by a filter located between the LNA and the mixer.

Chapter 3

A $\Delta\Sigma$ Frequency Synthesizer

In multichannel radio receivers channel selection can be achieved using a frequency synthesizer. Work by Jacquet [JK90] indicates that a fractional division digital PLL, with digitally achieved jitter reduction, is the most suitable approach for monolithic implementation of frequency synthesis in order to achieve high resolution and low complexity.

Figure 3.1 : A $\Delta\Sigma$ fractional-N based PLL frequency synthesizer

The digital $\Delta\Sigma$ fractional-N PLL frequency synthesizer presented in this chapter is a printed circuit board version of the VLSI implementation currently being researched by Riley [RCK93]. Jitter reduction is achieved using a third order digital $\Delta\Sigma$ modulator to control the modulus division process.

3.1 A Fractional-N based PLL

The fractional-N based PLL component of figure 3.1 uses a dual modulus divider (DMD) and a modulus extension circuit (MEC) for prescaling the feedback signal [Zho91]. The average density of ones in the control bitstream driving the DMD provides the capability of generating output frequencies from $n(f_{ref})$ to $(n+1)f_{ref}$ in fractional steps. The resolution of this step is dependent upon the number of input bits on the $\Delta\Sigma$ modulator. By manipulating the density of ones using the MEC circuit the division capability is extended providing a range of approximately $n(f_{ref})$ to $(n+5)f_{ref}$. The generalized transfer function of a phase locked loop is given by

$$\frac{\theta_o(s)}{\theta_i(s)} = \frac{G(s)}{1 + G(s)H(s)} \quad (3.1)$$

where, from Figure 2.1

$$G(s) = (k_p)(F(s))(k_v/s) \quad (3.2)$$

and

$$H(s) = \frac{1}{n + \beta} \quad (3.3)$$

Here β can be any integer or fraction between 0.25 and 4.75 dependent upon the average density of ones in the control bit stream and the MEC circuit. This density is a function of the dual modulus divide ratio n , $n+1$, and the additional control bits C_0 and C_2 . The variable k_p is a tuning constant dependent upon the type of phase-frequency detector (PFD) employed, and is measured in (volt/rad). As well k_v , the tuning constant of the VCO, is measured in (rad/sec)/volt. The loop filter $F(s)$ consists of 4 effective stages of filtering with 2 stages generating dominant poles that affect the loop bandwidth.

The output of the PFD is initially low pass filtered using a passive RC structure, whose non-dominant pole (with respect to additional filtering incorporating dominant poles) breaks at approximately 5 times the loop bandwidth [Mot1]. The integrated signal is then processed by an active loop filter, traditionally used in second order PLLs. The equivalent single ended circuit diagram is shown in Figure 3.2 (b) [Mot1,UR83].

Figure 3.2 : a) Second order active loop filter with a compensating zero;
b) equivalent circuit of the filter of part (a).

The transfer function of the filter in Figure 3.2 (b) contains two poles and a compensating zero and is given by

$$F_1(s) = -\frac{1}{\tau_1 s} \times \frac{1 + \tau_2 s}{(1 + \tau_3 s)} \quad (3.4)$$

where

$$\begin{aligned} \tau_1 &= R_1 C_1 \\ \tau_2 &= R_2 (C_1 + C_2) \\ \tau_3 &= R_2 C_2 \end{aligned} \quad (3.5)$$

If only the first two stages of filtering are considered then the loop becomes a type 2 third order structure. The type 2 indicates there are 2 DC integrators, and the third order is indicated by the order of the denominator of the open loop transfer function below.

$$G_{(j\omega)} H_{(j\omega)} = -\frac{k_v k_p}{(n + \beta) \tau_1 \omega^2} \left[\frac{1 + j\omega \tau_2}{1 + j\omega \tau_3} \right] \quad (3.6)$$

This equation allows evaluation of both phase and magnitude of the open loop gain as a function of frequency.

Additional loop filtering, a requirement for suppression of reference frequency feedthrough, was provided using an active second order LPF. An emitter follower configuration, shown in Figure 3.3 (c), was chosen to ensure only a single phase inversion occurs around the loop [Mot1,UR83]. The emitter follower also provided a clamping action to the VCO tune voltage by constraining it to within V_{cc} and ground of the transistor rather than the +12 and -5 volts of the opamp.

Figure 3.3 : The complete type 2 fifth order loop filter

The addition of this filter creates a type 2 fifth order loop. From Figure 3.3 (c) we have

$$v_o = v_i - (i_1 + i_2)R_{ef} - i_1R_{ef} \quad (3.7)$$

$$v_o = v_i - i_1R_{ef} - i_2R_{ef} - i_2 \times \frac{1}{j4\omega C_{ef}} \quad (3.8)$$

assuming a gain of 1, and no input current to the transistor then additionally

$$v_o \cong i_1 \times \frac{1}{j\omega C_{ef}} \quad (3.9)$$

Rearranging and substituting (3.9) into (3.7) leads to

$$i_2 = \frac{v_i - v_o - 2j\omega R_{ef}C_{ef}v_o}{R_{ef}} \quad (3.10)$$

Substituting (3.10) and (3.9) into (3.8) and rearranging provides

$$F_2(j\omega) = \frac{1}{1 + 2j\omega R_{ef}C_{ef} - (2\omega R_{ef}C_{ef})^2} \quad (3.11)$$

This results in a Sallen-Key filter with a gain of 1, and from this second order form normalized to ω_n , the cutoff frequency for the emitter follower circuit is found to be

$$\omega_{c, ef} = \frac{1}{2R_{ef}C_{ef}} \quad (3.12)$$

An arbitrary choice of R_{ef} between 1 k Ω and 100 k Ω provides the basis for the design of this filter. The magnitude response is unity over the passband with a rolloff of 40 dB/decade after the cutoff frequency. The generalized open loop transfer function for the filter can also be described by

$$F_2(j\omega) = \frac{\omega_n^2}{-\omega^2 + j2\zeta\omega\omega_n + \omega_n^2} \quad (3.13)$$

where ω_n is the natural frequency and ζ the damping ratio of the filter. If equations (3.6) and (3.13) are multiplied, the complete open loop transfer function for the PLL becomes

$$G(j\omega)H(j\omega) = -\frac{\omega_n^2 k_v k_p}{(n + \beta)\tau_1 \omega^2} \left[\frac{1 + j\omega\tau_2}{j\omega [2\zeta\omega_n + \tau_3(\omega_n^2 - \omega^2)] + \omega_n^2 - \omega^2 - 2\zeta\tau_3\omega_n\omega^2} \right] \quad (3.14)$$

This equation can be used to evaluate the magnitude and phase of the open loop transfer function. The use of the emitter follower filter for the additional loop fil-

tering allows a design that uses two degrees of freedom. Both ω_n and ζ can be varied independantly for the realization of a stable loop filter. In the actual filter design equation (3.12) was used in the determination of the filter components.

The passive LPF of Figure 3.3 (b) was used to transform the low impedance output of the first stage of filtering to a slightly higher impedance level for better matching to the emitter follower stage. The LPF was also used to effectively isolate the ground planes between the two filter stages. Above the cutoff frequency most of the AC component was coupled to the ground plane of the first stage of filtering. The cutoff frequency of this isolating filter was set to roughly 10 times the loop bandwidth to ensure minimal degradation of loop filter phase margin. For this reason it has not been included in the transfer function of equation (3.14). The transfer function of the matching filter is given by

$$F_m(j\omega) = \frac{1}{1 + j\omega\tau_m} \quad (3.15)$$

where

$$\tau_m = R_m C_m \cong \frac{1}{10 \times BW_{loop}} \quad (3.16)$$

The equations of this section will be used for component design in a subsequent section that deals with the realization of the phase locked loop circuit.

3.1.1 The Division Factor

From the circuit of figure 2.1 the division components shown consists of a single dual modulus divider (DMD) and an additional modulus extenson circuit (MEC) which extends the divider capability from $(n/n+1)$ to $(n/n+5)$. This follows an approach used by Riley [Ri3] and Zhong [Zho91].

The modulus extension is divided into two functions. The first function is a fixed $\div 5$ circuit implemented as a Johnson counter. The second function is a shift register that multiplexes in two control signals and generates a five bit serial sequence that controls the divide ratio of the DMD. The provision of the two input control signals allows the capability of the extended division ratio. Figure 3.4 shows the functional diagram of the DMD with its additional control.

Figure 3.4 : The dual modulus divider with the modulus extension circuit.

From this figure LD and CLK_2 are 10MHz signals derived from the 50MHz input signal f_{in} of the $\div 5$ circuit. LD is used to load the shift register before the shift sequence begins. CLK_2 is used to register inputs C_0 , C_2 and $b(t)$. C_0 and C_2 are input bits that are multiplexed with $b(t)$ to produce the control signal $CNTRL$ which extends the division ratio of the DMD. The signal $b(t)$ is the single bit bistream output of the $\Delta\Sigma$ modulator. Figure 3.5 shows the block diagram of the Johnson counter $\div 5$ circuit.

Figure 3.5 : Johnson counter divide by 5 circuit.

Note : f_{in} common
to all flip flops

The design of the multiplexed shift register includes features to minimize interference between the bitstream and the divider output. The extended division control of the DMD is based upon a five bit bitstream of data. The design begins with an arbitrary selection of five bit sequences that link the density of ones to the divide

ratio.

Figure 3.6 shows the timing diagram that relates how bits D_0 through D_4 are developed into the $CNTRL$ signal from control signals C_0 , C_1 , C_2 , and $b(t)$. The term C_1 is removed through redundancy and $b(t)$ represents the output bitstream of the $\Delta\Sigma$ modulator, that develops into the extended control signal $CNTRL$ (see Figure 3.7). The rising edge of the DIV signal, which feeds the PFD, must be free of interfering noise and therefore no level changes are allowed to occur between bits D_2 and D_3 [MC90]. The density of low bits in D_0 through D_4 is directly related to the overall division factor of 80 through 85. For instance, all five bits high indicates a divide by 80, while one of five bits low indicates a divide by 81. The signal $b(t)$, the $\Delta\Sigma$ modulator output, is used to generate bits b_1 and b_2 as seen in both Table 3.1 and Figure 3.7. Yet additional control bits C_0 and C_2 are required to develop the full five bits required for extended division control. Signals C_0 and C_2 originally begin as C_0 , C_1 , and C_2 , yet redundancy in C_0 and C_1 allow C_1 to be eliminated from the design equations.

Figure 3.6 : Bit sequence relationships that control the divide ratio.

Table 3.1 shows the mapping of input to output bits. There are 32 possible states based on the five input bits. Only 12 of these states (sections A, B and C) are required to define the necessary output states that correspond to a designated division modulus. The resulting modulus extension from the mapping leads to a division range of 80 through 85, based upon the original 16/17 dual modulus divider.

From the mapping of Table 3.1 a set

TABLE 3.1 : Mapping of input to output bits for extended modulus control.

Section	N	C ₀	C ₁	C ₂	b ₁	b ₂	D ₀	D ₁	D ₂	D ₃	D ₄
A	80	0	0	0	0	0	1	1	1	1	1
	81	0	0	0	0	1	1	1	1	1	0
	81	0	0	0	1	0	1	1	1	1	0
	82	0	0	0	1	1	0	1	1	1	0
B	81	0	0	1	0	0	0	1	1	1	1
	82	0	0	1	0	1	0	0	1	1	1
	82	0	0	1	1	0	0	0	1	1	1
	83	0	0	1	1	1	0	0	1	1	0
C	83	1	1	1	0	0	1	0	0	0	1
	84	1	1	1	0	1	0	0	0	0	1
	84	1	1	1	1	0	0	0	0	0	1
	85	1	1	1	1	1	0	0	0	0	0

of equations for D_0 through D_4 can be generated using C_0 and C_2 to differentiate between sections A, B and C. Bits b_1 and b_2 generate the four states required to define the divide value of N for each section. The reduced Boolean equations, whose assignment is made on the rising edge of CLK_2 , are given by

$$D_0 = C_0 \cdot C_2 \cdot \bar{b}_1 \cdot \bar{b}_2 + \bar{C}_0 \cdot \bar{C}_2 \cdot \bar{b}_1 + \bar{C}_0 \cdot \bar{C}_2 \cdot \bar{b}_2 \quad (3.17)$$

$$D_1 = \bar{C}_0 \cdot \bar{C}_2 + \bar{C}_0 \cdot C_2 \cdot \bar{b}_1 \cdot \bar{b}_2 \quad (3.18)$$

$$D_2 = D_3 = \bar{C}_0 \cdot \bar{C}_2 + \bar{C}_0 \cdot C_2 \quad (3.19)$$

$$D_4 = \bar{C}_0 \cdot \bar{C}_2 \cdot \bar{b}_1 \cdot \bar{b}_2 + C_2 \cdot \bar{b}_1 + C_2 \cdot \bar{b}_2 \quad (3.20)$$

Equations (3.17) through (3.20) define the five serial output bits of Figure 3.6. These bits are loaded and shifted through a five bit shift register that is clocked by the signal f_{in} , as defined in Figure 3.4. This clock operates at five times the frequency of the clock driving the synchronized inputs of the multiplexer.

Figure 3.7 : Shift register with multiplexed inputs.

Equations Q_{B0} through Q_{B4} , whose assignment is true on the rising edge of f_{in} , define the output bits of the shift register and are given by

$$Q_{B0} = LD \cdot D_0 + \overline{LD} \cdot Q_{B1} \quad (3.21)$$

$$Q_{B1} = LD \cdot D_1 + \overline{LD} \cdot Q_{B2} \quad (3.22)$$

$$Q_{B2} = LD \cdot D_2 + \overline{LD} \cdot Q_{B3} \quad (3.23)$$

$$Q_{B3} = LD \cdot D_3 + \overline{LD} \cdot Q_{B4} \quad (3.24)$$

$$Q_{B4} = LD \cdot D_4 \quad (3.25)$$

In equations (3.21) through (3.25) LD , generated by the 5 Johnson counter, controls the bit loading and shifting in the shift register. The output of the shift register, Q_{B0} , is the bitstream signal $CNTL$ which drives the control pin on the DMD. Figure 3.7 shows the block diagram of the shift register circuit.

3.1.2 Circuit Realization

In the translation of any circuit design from paper to electronics, many aspects of layout and construction must be considered to effect a successful realization. Appendix A contains a set of wiring diagrams and a parts list for the construction of the fractional-N based PLL described in this chapter.

The PLL, a printed circuit board PCB design, was built around a differential I/O PFD that is ECL based. The differential connections provide some level of common mode suppression for power supply noise and/or common mode noise present on the device inputs. EMI filters and bypass capacitors were placed at all device power supply pins to further suppress power supply noise. Coupling of the reference frequency input to the PFD was accomplished using a 1:16 transformer to step up the impedance from 50 Ω and to convert to differential mode. This differential signal is then diode limited to provide ECL level swings. Finally the signal is AC coupled and added to an ECL DC bias from the PFD (V_{BB}) to provide the correct ECL input.

The differential outputs of the PFD drive the loop filter previously described in this chapter. The final section of this loop filter, the second order emitter follower circuit, was designed empirically to meet the phase stability requirements of the loop. The output signal of this circuit was scaled using a resistive voltage divider to provide attenuation of the loop gain, a requirement due to the large tuning constant k_v of the particular VCO (C-616, a standard component produced by Zcommunications [Zcom]). The output of the VCO is followed by a 10 dB resistive pad and amplifier section to provide isolation from unwanted signals and noise in the load circuits that might couple back into the VCO and modulate its output signal. The isolated output is then passed through a 3dB power splitter to drive an amplifier to provide a 1.8GHz PLL output, and to drive a $\div 2$ prescaler to cut the frequency in half. This signal is amplified again, and then passed through another 3dB power splitter to drive an amplifier to provide a 900 MHz PLL output, and to provide an AC coupled signal to the DMD input. This feedback signal is further divided through the extended divider circuit and then converted from a CMOS to an ECL level for comparison by the ECL PFD.

The $\Delta\Sigma$ modulator, working in conjunction with the modulus extension circuit, is connected to the control input of the DMD and determines the divide modulus per reference frequency cycle of the phase locked loop.

For each group of components a separate return to ground was used to minimize the creation of inductive ground loops [Mot2]. Microstrip transmission lines of approximately 50Ω were designed for the critical signal path from the output of the VCO to the input of the DMD. Each of the components on the critical path used 50Ω I/O in order to minimize the amount of mismatch along the signal path.

Table 3.2 lists the fifth order loop filter component values resulting from the set of design equations presented in section 3.1 of this thesis. The tuning constant of the phase detector, the measured tuning constant of VCO, and the loop bandwidth are shown below.

$$k_v = 2\pi \cdot 76 \text{ [(MHz) / V]}$$

$$k_\theta = \frac{0.8}{2\pi} \text{ [V / rad]}$$

$$BW_{3dB, ClosedLoop} = 80 \text{ kHz}$$

TABLE 3.2 : PLL loop filter components.

Resistance (k Ω)	Capacitance (pF)
$R_1 = 1$	$C_1 = 47$
$R_2 = 4.2$	$C_2 = 18000$
$R_{ef} = 10$	$C_{ef} = 27$
$R_m = 1$	$C_m = 1500$

3.2 A Third Order Digital $\Delta\Sigma$ Modulator

For the fractional-N PLL frequency synthesizer of Figure 3.1, the modulus input on the DMD requires a controlled bitstream density in order to provide an arbitrary output frequency resolution. A digital $\Delta\Sigma$ modulator, an interpolative coder, is chosen for the control signal because the spectrum of the quantization error exhibits a high pass characteristic. With appropriate antialias filtering and decimation

Figure 3.8 : Third order digital $\Delta\Sigma$ modulator with single bit bitstream output.

the low frequency component, a representation of the $\Delta\Sigma$ input signal, is transmitted while the high frequency component, primarily quantization noise, is suppressed.

The order of the coder is of fundamental importance because of differences in the shape of the output spectrum. Work by Candy & Benjamin [CB81] indicate that first order coders generate recurring patterns in the error signal, a function of input level, which result in spurious tones in the output spectrum. At low frequencies, these tones may interfere with baseband signals and lead to bit errors. The only advantage of lower order coders is their unconditional stability.

The output noise spectral density of higher order structures monotonically increases at greater rates per unit frequency (shifts quantization noise to higher frequencies), resulting in greater SNR or bit resolution per given baseband bandwidth. Another advantage of a higher order structure is the interdependence of order, resolution and oversampling ratio. For a given resolution, increasing the order allows a reduction in the oversampling ratio. A grave disadvantage of third order (and higher) cascaded feedback structures is that they are subject to instability. Design of suitable architectures tends to be empirical in nature due to the sensitivity of stability upon initial conditions and scaling factors. Work has been done by Riley [RC92a] and Pinault and Lopresti [PL92] to define the bounds of stability for second order coders.

In this thesis a third order digital architecture was chosen with 28 input bits and a single bit bitstream output. This effective input bit resolution results in an effective frequency synthesizer output resolution of fractions of a Hz. The requirement for 28 bits in and 1 bit out led to the development of a digital architecture based on the use of a MASH3 type structure [Lon88,WW91], followed by a third order cascaded feedback structure [RCK93] used for concentrating the multibit noise shaped signal to a single bit bitstream. Figure 3.8 shows a block diagram of the modulator used to control the division process of the PLL.

3.2.1 MASH3 Digital $\Delta\Sigma$ Structure

A MASH3 structure is, by definition, a third order $\Delta\Sigma$ modulator composed of three cascaded first order coders. The advantage of this architecture is that higher order structures are always stable due to the unconditional stability of the fundamental unit, the first order structures.

A Z domain analytical representation of a first order structure, derived by Longo [Lon88], is presented below.

$$v_o(z) = x(z) + (1 - z^{-1})Q(z) \quad (3.26)$$

Here $x(z)$ represents the input signal and $Q(z)$ represents the quantization noise. The product of the integrator gain α and the quantizer gain k here are assumed equal to 1, where the quantizer gain adjusts itself to produce unity loop gain. Equation (3.26) shows that the output of the modulator contains a representation of the input signal plus quantization noise. The multiplication factor preceding the quantizer $Q(z)$ is a highpass function indicating that the quantization noise is noise shaped through this feedback action. A definition for the high pass function is given by

$$|1 - z^{-1}| = 2 \sin\left(\frac{\pi f}{f_s}\right) \quad (3.27)$$

where

$$z = e^{j\omega T} \quad (3.28)$$

From equation (3.27) an increase in the sampling frequency f_s decreases the magnitude of the high pass function, leading to a greater signal-to-quantization-noise ratio $SQNR$.

If a third order structure consisting of three single stages cascaded, as shown in Figure 3.9, is considered, then from Longo [Lon88] the output signal from each stage is

$$D_1(z) = x(z) + (1 - z^{-1}) Q_1(z) \quad (3.29)$$

$$\begin{aligned} D_2(z) = & z^{-1} (\alpha_1 H_1 - F_1) x(z) + z^{-2} (F_1 - \alpha_1 H_1) Q_1(z) \\ & - z^{-1} F_1 \cdot Q_1(z) + (1 - z^{-1}) Q_2(z) \end{aligned} \quad (3.30)$$

$$\begin{aligned} D_3(z) = & z^{-2} (\alpha_1 H_1 - F_1) (\alpha_2 H_2 - F_2) x(z) \\ & - z^{-2} F_1 (\alpha_2 H_2 - F_2) Q_1(z) \\ & + z^{-3} (F_1 - \alpha_1 H_1) (\alpha_2 H_2 - F_2) Q_1(z) - z^{-1} F_2 Q_2(z) \\ & + z^{-2} (F_2 - \alpha_2 H_2) Q_2(z) + (1 - z^{-1}) Q_3(z) \end{aligned} \quad (3.31)$$

Here α_i and k_i are the integration and quantizer gains respectively. The term H_i is the digital gain factor per stage while F_i is the feedback scaling factor. Given that $D_0(z)$ is the sum of the outputs of the three stages after appropriate delay processing, and if it is assumed that $H_1 = 1/\alpha_1$, $H_2 = 1/\alpha_2$, $F_1 = 1$, $F_2 = 1$, then $D_0(z)$ is given by

$$D_o(z) = z^{-2} x(z) + (1 - z^{-1})^3 Q_3(z) \quad (3.32)$$

This result clearly indicates that third order quantization noise shaping is present in the output signal. The output also contains a representation of the input signal pro

Figure 3.9 : A third order (MASH) $\Delta\Sigma$ modulator.

cessed by two delays.

In order to convert Longo's model from the sampled data model presented in Figure 3.9 to the digital accumulator model of Figure 3.8 it is necessary to assume that each of the $\Delta\Sigma$ stages of the modulator can be replaced by a N-bit digital phase accumulator (DPA) and that $D_1(z)$, $D_2(z)$ and $D_3(z)$ represent the carry outputs of each stage. Before proceeding to describe the third order structure used a comparison between a first order $\Delta\Sigma$ modulator and a first order DPA is presented to verify the above assumption. These results are based on work by Riley and Copeland [RC92b]. Consider the diagram of Figure 3.10 (a). The difference equation for such a DPA is given by

$$R_n = R_{(n-1)} + K_{n-1} - M \cdot \text{carry} [R_{(n-1)} + K_{n-1}] \quad (3.33)$$

Here R_n is the remainder in the accumulator at the end of cycle n , and K_n is the input value added to the accumulator during cycle n . The term M represents the carry output value of the DPA and is given by 2^k , where k is the number of bits in the accumulator. The carry function is given by

$$\text{carry} = R_{n-1} + K_{n-1} - \frac{M}{2} \text{sign} (R_{n-1} + K_{n-1} - M) - \frac{M}{2} \quad (3.34)$$

If $(K_n - M)$ is added to both sides of (3.33) then

$$(R_n + K_n - M) = (R_{n-1} + K_{n-1} - M) + K_n - \frac{M}{2} - \frac{M}{2} \text{sign}(R_{n-1} + K_{n-1} - M) \quad (3.35)$$

and if x_n and u_n are defined as

$$\begin{pmatrix} x_n \equiv R_n + K_n - M \\ u_n \equiv K_n - \frac{M}{2} \end{pmatrix} \quad (3.36)$$

then (3.33) can be rewritten in the form

$$x_n = x_{n-1} + u_n - \frac{M}{2} \text{sign}(x_{n-1}) \quad (3.37)$$

This equation can be mapped to the block diagram of Figure 3.10 (b), which is that of a first order $\Delta\Sigma$ modulator.

Figure 3.10 : a) a first order DPA b) a first order $\Delta\Sigma$ modulator

If the third order modulator of Figure 3.9 is now reconsidered, then each of the

first order coders can be replaced by 25 bit DPAs. This provides a 25 bit third order DPA with carry signals defined as $D_1(z)$, $D_2(z)$ and $D_3(z)$. The transformation process from $D_1(z)$, $D_2(z)$ and $D_3(z)$ to $D_0(z)$ of Figure 3.9 is accomplished using the linear transformation function $F_7(z)$ and $D2B$, shown in Figures 3.11 and 3.12.

Figure 3.11 : $F_7(z)$, the third order quantization noise shaping function.

This function processes the 3 DPA carry outputs with pipelining, defined in Figure 3.9, to provide a binary output. The number of 1's in the output of $D2B$ represent $D_0(z)$. The delay functions of Figure 3.9 can be expanded as follows.

$$(1 - z^{-1})^2 = 1 - 2z^{-1} + z^{-2}$$

$$z^{-1}(1 - z)^{-1} = z^{-1} - z^{-2}$$

The output of $F_7(z)$ is processed by function $D2B$ as shown in Figure 3.8. This function inputs the 7 bits and counts the number of high bits on each input cycle and represents them as a binary two's complement 4 bit output. Figure 3.12 shows a block diagram of the function. The binary subtract 3 circuit is used to adjust the magnitude of the output bits of the $D2B$ circuit such that this magnitude is centered around the input range of the concentrator circuit.

Figure 3.12 : Denury to binary (D2B) converter circuit

3.2.2 A Third Order $\Delta\Sigma$ Concentrator

The third order $\Delta\Sigma$ modulator presented in this section concentrates the 4 bit output of the D2B circuit of Figure 3.12 to a single bit bitstream, as indicated in Figure 3.8. This work, originally by Chao et. al. [CNL90], was empirically modified by Riley [RCK93] to develop the correct digital feedforward scaling factors. Figure 3.13 shows the modified block diagram of the modulator by Riley where the feedback factor M is implemented as a sign bit on the 4 bit adder.

The linearized transfer function for this modulator is given below, where k is assigned a value of $1/4$ to fit the measured spectrum.

$$H(z) = \frac{(1 - z^{-1})^3}{(1 - z^{-1})^3 + k \left(1 + 4(1 - z^{-1}) + 8(1 - z^{-1})^2 \right)} \quad (3.38)$$

The theoretical power spectral density of this third order digital $\Delta\Sigma$ modulator has been presented by both Chao and Riley and therefore the exercise is not repeated here.

The feedforward gain factors 1, 4, and 8 were implemented by bit shifts between the output bits of each respective accumulator and the feedforward input bits of the next adder stage. Each of these gain components differed by some power of 2 and

Figure 3.13 : A third order $\Delta\Sigma$ modulator with single bit bitstream output.

were therefore easily implemented in the digital modulator.

3.2.3 Circuit Realization

The full $\Delta\Sigma$ modulator of Figure 3.8 was implemented using a single field programmable gate array (FPGA) by Xilinx [Xil92]. The LCA 3090 was chosen because the device resources provided the number of configurable logic blocks (CLBs) required to accommodate the entire digital design. A printed circuit board was designed, fabricated, and then populated with a single LCA, an EPROM and other related hardware. This finished version was capable of providing third order quantization noise shaping to an operating frequency of 14 MHz.

Figure 3.14 shows the functional simulated power spectral density of the bistream output for the digital modulator circuit using a 10 MHz reference frequency. This simulation was accomplished by generating a netlist of the schematic file and converting it to a format compatible with PC Silos.

A complete set of logic and wiring diagrams for the digital modulator can be found in Appendix B.

3.2.4 Circuit Measurements

The power spectral density of the quantization noise was measured by capturing the output bit stream of the modulator for a fixed digital (DC) input. The bitstream

Figure 3.14 : Simulated power spectral density of the complete third order digital $\Delta\Sigma$ modulator

was captured using a computer card developed specifically for the purpose of grabbing bits. This bit grabber, synchronized to the 10 MHz clock driving the $\Delta\Sigma$ modulator, captured the bits and stored them in a RAM buffer to be subsequently transferred to a file in permanent memory on the computer. The bitstream file was then processed using software written by Riley [Ri2] to perform a DFT to generate and scale the power spectral density. Figure 3.15 shows the result of the measurement. The slope of the quantization noise is 3rd order which validates correct operation of the modulator. This result also compares favorably with the plot of the noise shaping of the power spectral density of the functional simulation shown in Figure 3.14.

3.3 Frequency Synthesizer Measurement

In this section, several fundamental measurements are presented that indicate the quality of the frequency synthesizer of Figure 3.1. These measurements include phase noise, transient response, and minimum frequency resolution.

Figure 3.15 : Measured power spectral density of the complete third order digital $\Delta\Sigma$ modulator

3.3.1 Phase Noise Response

Phase noise is a random short term frequency fluctuation which results generally from phase modulation of a carrier [HP1]. These short term instabilities in phase result in the creation of continuous noise sidebands which stretch out symmetrically from the carrier.

In radio receivers the level of the phase noise in the LO sidebands is translated directly to the IF sidebands. For a multichannel environment, sensitivity to a strong signal in an adjacent channel is dependent upon the level of phase noise at a full channel spacing from the carrier. In a frequency modulated system the presence of high levels of phase noise within the channel leads to reduced GNR for analog systems, and increased bit error rates for digital systems.

In a PLL frequency synthesizer the contour of the noise spectrum is dependant not just on the source reference but also on each of the components comprising the loop. Sorfleet [Sor91] presents a pertinent analysis of PLL synthesizer phase noise but limits the analysis to three noise sources, the reference frequency, VCO, and $\Delta\Sigma$ modulator. Poole [Po90] deals with synthesizer noise from a qualitative perspec

Figure 3.16 : Single sideband phase noise response of an integer division frequency synthesizer at 800 MHz.

tive and provides a strong descriptive work on a more complete set of noise sources in a PLL frequency synthesizer. Sources of noise that can also contribute to the phase noise response are the phase-frequency detector, the frequency divider, and the loop response.

The multiplication factor of the divider (N) degrades the phase noise response of the loop by adding $20\log(N)$ [dB] directly to the phase contour with respect to the VCO output. In effect the noise of the frequency reference, the PFD, and the loop filter are all multiplied by a factor of N .

The frequency reference allows some tuning of the phase contour, but acceptable levels of source jitter are dictated by the application. These levels can be manipulated by replacement with a better quality component.

The digital phase detector and active loop filter are both capable of adding noise to the phase contour. The PFD becomes a noise generator when the loop is in lock and the detector is operating in the dead zone. In this state the device outputs pulses that tend towards a zero width and therefore produce a detector gain that tends to zero. Noise can be produced with no change in the tune voltage if large detector gain fluctuations occur around the zero gain state. This problem can be remedied by introducing a DC phase offset into one of the differential inputs of the

Figure 3.17 : Single sideband phase noise response of a fractional division frequency synthesizer at 810 MHz.

loop filter. This offset leads to a locked condition in the loop for a given input phase offset resulting in output pulses that can range from zero to 360 degrees width. Noise in a PFD can also result from insufficient power supply decoupling. Proper selection of capacitance across the power supply terminals of the PFD component will help to reduce degradation in the noise performance of the PFD.

The loop response contributes to the phase contour in two ways, bandwidth and loop stability. The loop bandwidth is a design variable that must be chosen critically to ensure an optimum overall response between in-band suppression and out-of-band unsuppressed noise components (primarily VCO noise). Loop stability is ensured by designing the loop filter such that the loop gain falls off at 20 dB/decade at the unity gain crossover point.

VCO noise can arise from a number of different sources. Close in noise levels are a function of both the Q of the component and flicker or 1/f noise. Flicker noise can be reduced to some extent in bipolar circuits by using an emitter degeneration resistance to provide low frequency feedback. The Q of the VCO is a parameter that is directly associated with the tuning constant of the component. Inherently, VCOs with large tuning range will have lower Q values and therefore higher close in phase noise. At further out frequencies noise levels are more dependant upon noise figure (i.e. thermal noise), which indicates that noise levels will be depen-

dant upon the power levels and type of device used in the construction of the VCO.

What is presented here is the actual phase noise performance measurements of the synthesizer operating in both integer and fractional-N states. Figures 3.16 and 3.17 show the single sideband phase noise measurements respectively using Hewlett Packard's HP3048A phase noise measurement system [HP2] at 800 MHz and 810 MHz.

3.3.2 Transient Response

The transient response of the synthesizer was measured in frequency for both integer and fractional frequency steps, as shown in Figures 3.18 and 3.19.

Figure 3.18 : Transient response for an integer frequency step of 20 MHz.

The measurements, based on the use of the Hewlett Packard HP5372A Time Interval Analyzer [HP3], follow a procedure similar to that used by Sorfleet [Sor91]. The results are suitable for a rough indication of the loop dynamics but are limited by the measurement resolution of the HP5372A. Effects of the sampling noise by the HP5372A can be seen in Figures 3.18 and 3.19, and is measured to be approximately 300 KHz peak to peak.

Intended uses of the synthesizer in either broadband coded FM (slithering) or fre

Figure 3.19 : Transient response for a fractional frequency step of 5 MHz.

quency hopping receivers dictates the need for both large (integer) and small (fractional- n) steps in the transient analysis. The integer step in frequency indicates the worst case scenario for loop response in a large hop in frequency. The fractional step in frequency will indicate the quality of the loop dynamics while operating with an average division ratio and a high level of quantization noise.

3.3.3 Minimum Frequency Resolution

The frequency synthesizer is controlled by a $\Delta\Sigma$ modulator with 28 bits on the input. The synthesizer output is capable of a maximum jump in frequency (f_{jump}) equal to β_{max} times the PLL source reference, f_{ref} when the prescaler is switched from n to $(n + \beta_{max})$. This means the synthesizer will jump a total average frequency of $(10 \times \beta_{max})$ MHz when the DMD (SP8782A) input is changed from n to $(n + \beta_{max})$. If $\beta_{max} = 1$ then the synthesizer will jump 10 MHz.

The minimum frequency resolution of the synthesizer is determined by measuring the difference in frequency when the input bits on the $\Delta\Sigma$ modulator are toggled by a LSB. The minimum frequency resolution can be determined using

$$f_{res} = \frac{f_{jump}}{2^N} \quad (3.39)$$

Here N is the number of input bits on the $\Delta\Sigma$ modulator. Equation (3.39) indicates an expected minimum frequency resolution of 0.0373 Hz. The actual measured minimum resolution was found to be approximately 1 Hz for a movement of the 5th significant bit. This correctly corresponds to the theoretical value of $0.0373 \cdot 2^5 \sim 1.19$ Hz. The frequency counter used for measuring this resolution was incapable of measuring less than 1 Hz for an input of 800 MHz.

Chapter 4

A $\Delta\Sigma$ Frequency Discriminator

In digital FM radio receivers analog to digital conversion and data recovery can be achieved using a frequency discriminator. The discriminator presented in this chapter is a printed circuit board version of the VLSI $\Delta\Sigma$ based frequency discriminator proposed by Beards and Copeland [BC92].

In general, a frequency discriminator provides an output signal proportional to the frequency deviation of the input [ZT85]. For an ideal discriminator, if the input signal is given by:

$$x(t) = A_c \cos [\omega_c t + \phi(t)] \quad (4.1)$$

The output will be

$$y(t) = \frac{1}{2\pi} K_D \frac{d\phi}{dt} \quad (4.2)$$

For an FM system $\phi(t)$ is given by

$$\phi(t) = 2\pi f_d \int_0^t m(\alpha) d\alpha \quad (4.3)$$

This allows (4.2) to be rewritten as

$$y(t) = K_D f_d m(t) \quad (4.4)$$

Here K_D is the discriminator constant (*volts/hertz*) and $y(t)$, the discriminator output. Under ideal operation the discriminator will have a linear frequency to voltage transfer function with an output magnitude of zero for $f=f_c$.

The $\Delta\Sigma$ frequency discriminator described in this thesis, based on the work of Beards and Copeland [BC92], provides a digital output whose density of ones $\bar{B}$ is proportional to frequency deviations of the input signal. The density of ones is given by

$$\bar{B} = -K_D f_d(t) \quad (4.5)$$

For an $n/(n+1)$ DMD and a bipolar $\pm V_{DD}/2$ output, then

$$K_D = \frac{V_{DD}/2}{f_{ref}/2} = \frac{V_{DD}}{f_{ref}} \quad (4.6)$$

and

$$f_d(t) = f_{in} - \left(n + \frac{1}{2}\right) f_{ref} \quad (4.7)$$

where $nf_{ref} < f_{in} < (n+1)f_{ref}$.

Here f_{in} is the frequency modulated input and n is the lower moduli of the $n/(n+1)$ divide process.

For a digital bipolar system the discriminator output signal is given by

$$V_O(t) = -K_D f_d(t) + V_Q(t) \quad (4.8)$$

which is similar in form to the equation for an ideal discriminator. The negative sign indicates that an increase in frequency leads to a decrease in bit density, which can be arbitrarily removed by using $\bar{Q}$ on the output of the quantizer thus changing the definition of K_D from a positive to a negative result. The component $V_Q(t)$ is the quantization noise associated with the oversampling $\Delta\Sigma$ operation.

In the process of converting frequency deviation information into a digital bit-

stream, quantization noise is produced that has a power spectral density characterized by a high pass transfer function. Post filtering must be applied to remove this noise component and then decimation applied for reduction to baseband rate.

4.1 Circuit Operation

Consider the block diagram of Figure 4.1(a), a simplified model for a first order $\Delta\Sigma$ discriminator.

(a)

(b)

Figure 4.1 : (a) Equivalent block diagram of a first order $\Delta\Sigma$ frequency discriminator; (b) modified block diagram of a first order $\Delta\Sigma$ frequency discriminator (reprinted from [BC92].)

The discriminator feedback factor is dependant upon the magnitude of the average frequency of the input signal. In the classic sense of $\Delta\Sigma$ modulation this dependency is not present, yet if narrow band modulation is assumed, then the small per-

cent deviation δ on the input can be neglected leading to a definition for the coarse quantization time edge deviation output of

$$q = \frac{T_{in,k}}{2} = \frac{T_{carrier}}{2} (1 - \delta + \delta^2 - \dots) \cong \frac{T_{carrier}}{2} \quad (4.9)$$

For small deviations on f_{in} the feedback signal q is a time edge movement approximately equal to $1/2$ of the carrier input period. The useful information contained in the input signal is the deviation due to the modulation of the carrier. An approximate model is shown in Figure 4.1 (b), where the input signal is redefined as

$$\bar{T}_{in,k} = [(n + \beta) T_{in,k} - T_{ref}] \quad (4.10)$$

Here n is an integer and β a fraction between 0 and 1, and the sum of the two defines the average division factor on f_{in} . The valid range of operation for (4.10) is $nT_{in,k} < T_{ref} < (n+1)T_{in,k}$.

Consider the simplified model of Figure 4.2, presented by Beards and Copeland [BC92], for the second order $\Delta\Sigma$ frequency discriminator.

Figure 4.2 : Model for the second order $\Delta\Sigma$ frequency discriminator (reprinted from [BC92].)

The difference equations defining the operation of this model are given by

$$v_{c,k} = v_{c,k-1} + \alpha \tau_k - \text{sgn}(v_{c,k-1}) v_{r2} \quad (4.11)$$

(a)

(b)

Figure 4.3 : a) Block diagram of the second order $\Delta\Sigma$ frequency discriminator;
 b) waveforms describing correct discriminator operation
 (reprinted from [BC92].)

$$\tau_k = \tau_{k-1} + [(n + \beta) T_{in, k-1} - T_{ref}] - \text{sgn}(v_{c, k-1}) \frac{T_{in, k-1}}{2} \quad (4.12)$$

Here α is the gain constant between the cascaded stages, and v_{r1} and v_{r2} the gain factors on the magnitude of the feedback signals from the first and second stages respectively.

The $\Delta\Sigma$ frequency discriminator of Figure 4.3 (a) is the discriminator circuit implemented in this thesis. For this second order structure the waveforms of Figure 4.3(b) and the definitions of a first order structure provided in equations (4.9) and (4.10) allow for equation (4.12) to be rewritten as

$$\tau_k = \tau_{k-1} + \bar{T}_{in,k-1} - \text{sgn}(v_{c,k-1}) \frac{T_{carrier}}{2} \quad (4.13)$$

Equations (4.11) and (4.13) will remain valid as long as the oversampling rate f_{ref} is much greater than the rate of deviation on the input signal (modulation rate) and as long as $nf_{ref} < f_{in} < (n+1)f_{ref}$.

From a qualitative perspective the circuit of Figure 4.3 (a) operates by comparing f_{out} , the divided down signal of f_{in} , with f_{ref} . Depending on whether f_{out} is phase advanced or retarded with respect to f_{ref} , the output of the PFD will switch charge pump #1 to either pump current on, or draw current off the integrating capacitor. On each cycle of $T_{ref} = (1/f_{ref})$ voltage v_c is compared against a known reference level and the output is quantized.

This output is a one bit approximation of the input phase difference $\bar{T}_{in,k-1}$, and is used to control the switching of charge pump #2 to provide second order statistics, as well as toggle the modulus control of the DMD to reduce the phase error between f_{out} and f_{ref} .

4.2 Circuit Design

For this thesis, the design of the $\Delta\Sigma$ discriminator circuit was based upon the use of off-the-shelf components. In this section each component of the design shown in Figure 4.3 is investigated to present its underlying operation and how it must be implemented in order to ensure correct feedback sense and to ensure that an entire discriminator cycle occurs within a single period of the reference clock.

The design parameters for the discriminator were arbitrarily chosen to be

$$f_{in} = 100 \rightarrow 110\text{MHz}$$

$$\text{Deviation} = 100\text{KHz}$$

$$\text{Modulation} = 50\text{KHz}$$

$$f_{ref} = 10\text{MHz}$$

Here *Deviation* and *Modulation* are the frequency deviation and the modulation rate on f_{in} . The relationship between f_{ref} and the f_{in} range is dependant upon the value and ratio of moduli on the DMD. In this case n equals 10, and the DMD toggles between 10/11.

4.2.1 The Dual Modulus Divider (DMD)

The divider chosen for the discriminator was the Plessey SP8680B, a 600MHz $\pm 10/11$ device [Ples]. Internally the device is constructed of a 3 stage ECL Johnson counter with pulse swallowing techniques and a divide by 2 output stage to provide the 10/11 divide capability.

It has a single ended input with an internal 50K Ω pulldown resistor, and a differential output which requires external pulldown resistors to -5.2 volts.

Figure 4.4 : Timing diagram for the Plessey SP8680, (reprinted from [PLES]).

From Figure 4.4 the set up time t_s is defined as the minimum time that can elapse between a $L \rightarrow H$ transition of the control input and the next $L \rightarrow H$ input clock pulse

transition to ensure that the $\div 10$ mode is obtained. As well, the release time t_r is defined as the minimum time that can elapse between a $H \rightarrow L$ transition of the control input and the next $L \rightarrow H$ clock pulse transition to ensure that the $\div 11$ mode is obtained.

4.2.2 The Phase Frequency Detector (PFD)

The digital phase frequency detector chosen for the discriminator circuit was the Sony SPECL CXB111ZQ [Son89]. This is a fully differential ECL component that accepts data rates to 800MHz. The very short rise, fall and propagation times of this high speed device keep the width of the nonlinear distortion region (dead zone) reasonably small. In a closed loop structure such as the frequency discriminator, nonlinear distortion in the phase/voltage transfer function can lead to unwanted beat frequencies and noise translated into the baseband. The differential inputs on the device allow for suppression of common mode signals coupled to the input lines. As well the differential operation allows for suppression of power supply noise.

The operation of the PFD is summarized in the timing diagram of Figure 4.5.

Figure 4.5 : Timing diagram for the Sony PFD, CXB1112Q (reprinted from [Son89]).

If the rising edge of the reference input R leads the rising edge of the variable input V , then the up output U goes high on the rising edge R and resets on the next rising edge of V . As long as R leads V the pulse width of U will vary linearly with the phase difference of R and V , but the down output D will remain low. Similarly if the rising edge of V leads the rising edge of R , then on the rising edge of V the output D goes high and then resets on the next rising edge of R . As long as V leads R

then the pulse width of D will vary linearly with the phase difference of V and R , but the up output U remains low. If the timing edges of both V and R are synchronized then both D and U remain low producing a neutral state output. This description provides a way for understanding how the charge pump will work.

4.2.3 Charge Pump Circuit Operation

The charge pump circuit of the frequency discriminator was designed to convert the analog time edge information on the PFD differential outputs to analog voltage levels by switching quantized packets of current onto integrating capacitors. This provided the second order integration for the discriminator. The feedforward packets of current are a direct function of the pulse widths of the U/D signals and therefore linearly related to the time edge difference between the modulation and the reference signals on the input of the PFD.

The full charge pump circuit consists of a set of differential feedforward current switches, a set of differential feedback current switches, and a common mode feedback component to suppress common mode drift on the differential integrating capacitors.

The block diagram of Figure 4.6 indicates the simplified operation of a set of differential current switches.

Figure 4.6 : Simplified differential charge pump

levels on C_1 and C_2 through sense resistors R_{S1} and R_{S2} . If C_1 and C_2 are charging differentially then the voltage level seen on the base of Q_1 and Q_2 remains unchanged and currents I_{cm1} and I_{cm2} remain static. If the charge on capacitors C_1 and C_2 begin to drift up, resistors R_{S1} and R_{S2} develop a rising voltage on the base of Q_1 and Q_2 . This level is compared to the reference level across R_b , where the difference can lead to Q_1 and Q_2 turning on more deeply drawing more current off of C_1 and C_2 bringing their levels down to the reference level set by the base of Q_3 and Q_4 . Alternatively, if C_1 and C_2 drift toward $-V$ then by the same negative feedback mechanism Q_1 and Q_2 can turn on less deeply drawing less current off C_1 and C_2 allowing them to charge up to the reference level developed across R_b .

4.2.3.1 Charge Pump Circuit Design

The design of the charge pump circuit is based upon some arbitrary choices about initial design parameters. From Figure 4.3 (a) the full voltage (i.e. dynamic) range on the integrating capacitor was set to

$$\Delta V_{TOT} = \frac{10 \times \Delta Q}{C} \quad (4.14)$$

The discrete charge component ΔQ is based upon a step size of $1/4$ the input period T_{in} , and is given by

$$\Delta Q = \frac{T_{in}}{2} \times I_1 \quad (4.15)$$

The discrete charge component contributed by the feedback current I_2 is a function of T_{ref} and is given by

$$\Delta Q_2 = \frac{T_{ref}}{4} \times I_2 \quad (4.16)$$

From Figure 4.3 (b) the effect of feedback current I_2 on the comparator voltage V_c is a fixed voltage step of

$$\Delta V_{C2} = \frac{\mp \Delta Q_2}{C} \quad (4.17)$$

for each cycle of f_{ref} . The polarity of the charge is dependant upon the instantaneous magnitude of $b(t)$.

The discrete feedforward charge component contributed by I_1 is dependant upon the difference between the falling edges of f_{ref} and f_{out} . Empirical results [R11] indicate that this difference τ_k is unlikely to exceed ten times the same polarity step size of $T_{IN}/2$. Typically τ_k will change polarity every 3 to 4 cycles due to the nature of the coarse quantization involved. The feedforward charge component is given by

$$\Delta Q_1 = \tau_k \times I_1 \quad (4.18)$$

with the effective change on V_C due to I_1 being

$$\Delta V_1 = \frac{\mp \Delta Q_1}{C} \quad (4.19)$$

The sign on ΔV_{C1} is dependant upon whether f_{out} is phase advanced (-ve) or phase retarded (+ve) with respect to f_{ref} .

The models for analog $\Delta\Sigma$ modulators typically symbolize the feedforward and feedback scaling factors by α_1 and α_2 respectively. In the charge pump circuit these scaling factors are directly a function of the current levels I_1 and I_2 .

The relationship $\alpha_1 = \alpha_2$ can be paralleled in the charge pump by setting $\Delta Q_1 = \Delta Q_2$. This relationship has been shown to provide stable operation [CT1] and can be expressed as

$$\frac{I_1 \times T_{in}/2}{C} = \frac{I_2 \times T_{ref}/4}{C} \quad (4.20)$$

Through the operation of the DMD, it is also known that

$$T_{in} = \frac{T_{ref}}{n + 0.5} \quad (4.21)$$

for a midrange fixed frequency input. Substituting (4.21) into (4.20) and rearranging yields

$$\frac{I_2}{I_1} = \frac{4}{2(n + 0.5)} \quad (4.22)$$

The application of the dynamic range $DR=10\Delta Q$ provides an additional design equation of

$$\frac{I_2 \times T_{ref}/4}{C} = \frac{DR/2}{10} \quad (4.23)$$

If Figure 4.8, a more complete model for the charge pump circuit, is considered then the above equations can be used to determine the required circuit elements.

Due to the large number of variables in the circuit design, initially some variables had to be chosen arbitrarily. Capacitors C_1 and C_2 were set to be $15pf$. Any value less than this would be within an order of magnitude of the parasitic capacitances of the charge pump transistors. The DMD chosen was to be a $\pm 10/11$ component. This allowed f_{in} to range from $100 \rightarrow 110MHz$ for a $10 MHz$ reference frequency. The dynamic range on C_1 and C_2 was set to 2 volts to match the full input swing on the latched comparator/quantizer. This allowed a ± 1 volt swing around the common mode bias point of -0.7 volts. From (4.23) I_2 was calculated to be $60 \mu A$, and substituting into (4.22) I_1 was found to be $315 \mu A$.

The common mode feedback circuit was designed interactively using Spice [Spi] simulation techniques based upon the initial assumption that

$$\frac{gm}{2\pi C} \cong UGBW \quad (4.24)$$

where

Figure 4.8 : Charge pump circuit with feedforward and feedback switching networks, including the common mode feedback circuit.

$$gm = \frac{I_{c,cm}}{V_T} \quad (4.25)$$

From the transconductance, the common mode current is given by

$$I_{c,cm} = V_T \times 2\pi C \times UGBW \quad (4.26)$$

For the transistors used (2N3904), with a unity gain bandwidth of $\cong 300\text{MHz}$ and a $V_T = kT/q = 26\text{ mV}$ (assuming a temperature of 300K), a common mode current of $735\mu\text{A}$ was determined. Yet from (4.25) the gain of the common mode feedback circuit is directly related to $I_{c,cm}$. To ensure a suitable amount of common mode suppression $I_{c,cm}$ was arbitrarily set to 4mA to raise the gain as much as possible. Open loop SPICE simulations of the charge pump circuit indicated a UGBW of 26MHz and a phase margin of 47.2 degrees . At 10 MHz , the frequency of operation for the comparator, the gain was 14.3dB . With $I_{c,cm}$ set to 4mA , the current sink for the common mode circuit $I_{c,sink}$ becomes 16mA .

Knowing the values of all other currents in the circuit, the single ended source currents were calculated using

$$I_{source} = I_1 + I_2 + I_{c,cm} \quad (4.27)$$

From (4.27) I_{source} was determined to be 4.375mA . Figure 4.9 (a) shows a diagram of the circuit used for the generation of the source currents. The 200Ω potentiometer used in the source current circuit allowed the introduction of a small DC offset current between each of the source currents. This in turn allowed the tuning of a fixed offset voltage between the levels on the differential charging capacitors. The offset provided the capability of steering the operation of the PFD away from the dead zone.

Figure 4.9 (b) shows the circuit diagram of a Widlar current source used for the realization of currents I_1 , I_2 , I_{sense} , and $I_{cm,sink}$.

For the sensing node of Figure 4.8 the sink current I_{sense} was arbitrarily set to 4mA . For the same circuit an emitter follower and diode connected transistor were used to isolate and level shift each of the sense node resistors R_{S1} and R_{S2} from the inte

Figure 4.9 : a) Source current circuit; b) current sink circuit.

grating capacitors.

A complete circuit diagram of the charge pump of Figure 4.8, including calculated values for bias components, can be seen in Appendix C.

4.2.4 The Comparator/Quantizer and Interface Circuit

The quantizer used in the discriminator is a high speed ECL pre-production device (MC10E1651) provided by Motorola [Mot3]. The I.C. consists of a pair of differential amplifiers with output latches. When the latch clock is high Q_{n+1} reflects the instantaneous result of the input comparison as an ECL output level. When the latch clock goes low the instantaneous result of the input comparison is latched as an output ECL level. In order to realize a complete flip flop action, the dual latched comparators were connected in a master-slave configuration.

Figure 4.10 shows a block diagram of the full discriminator circuit. This diagram shows an ECL PAL used to generate both reference and comparator clocks for the circuit.

The ECL PAL uses a ÷4 Johnson counter for the reference (R , $\bar{R}$) and comparator (CLO) clocks. The PAL is also used to generate a set of differential signals U_2 , $\bar{U}_2$, D_2 and $\bar{D}_2$ that drive the feedback current switches of the charge pump. As shown

Figure 4.10 : Block diagram of the complete discriminator.

in Figure 4.3 (b) current I_2 has a fixed period and duty cycle. In the design parameters the duty cycle was set at 25% or a pulse width of $T_{ref}/4$. For this reason a 40MHZ reference clock was used to generate the 10MHZ signals needed for the reference on the PFD. The two orthogonal outputs of the Johnson counter allowed

Figure 4.11 : The reference and feedback signal ECL PAL circuit.

for the derivation of a 10MHZ signal with the 25% duty cycle. This pulse was used to generate charge pump feedback signals. Figure 4.11 shows the diagram for the ECL PAL interface circuit.

4.3 $\Delta\Sigma$ Frequency Discriminator Circuit Realization

What is presented in this section is a set of measurements and analysis qualifying the performance of the $\Delta\Sigma$ frequency discriminator circuit shown in Appendix C. The circuit was realized as a two layer printed circuit board using small outline surface mount components.

Figure 4.12 shows the timing diagram for circuit operation under an FM input of 100 KHz deviation and 70 KHz modulation rate on a 107MHz carrier. From the measurement of Figure 4.12, five continuous points of operation have been identified for a single reference cycle. Point 1 indicates the rising edge of the reference signal $\bar{R}$, which is used as a clock to latch the comparator/quantizer element. The change in the output of the quantizer, Q , seen at point 2 occurs on the rising edge of $\bar{R}$.

Point 3 indicates the rising edge of $\bar{V}$, the output of the DMD. The difference between the rising edges of $\bar{R}$ and $\bar{V}$ contain a DC phase offset component and a modulation component τ_k which result in the pulse width modulation of the PFD outputs U_1 and D_1 .

Point 4 indicates the position for the rising edge of the 25% duty cycle feedback signals U_2 and D_2 . It is critical that the falling edge of these signals occur before the next rising edge of $\bar{R}$ seen at point 5. This measurement indicates that a full cycle of the second order structure occurs within one period of the reference frequency, and therefore occur with a single delay around the loop. This criterion is critical for stable operation of a feedback $\Delta\Sigma$ structure. Measurements also indicated that the useful tuning range for the DC phase offset was 50% of the reference period. Noise shaping operation did not occur beyond this range. In the timing diagram, at point 3, the edge information τ_k on $\bar{V}$ contains the coarse quantization noise as well as the digital FM modulation signal.

Figure 4.13 presents a 120 set ensemble average of 16000 point FFTs on the output bitstream of the discriminator. A total of 2^{20} bits were captured and stored to allow a post processing average of 120 separate 16000 point FFTs. The measurement clearly exhibits quantization noise shaping. The 70KHz modulation tone (effectively a 140kb/s rate) is also present as expected. The signal power level of the tone appears to be about 28 dB above the noise level at 70KHz, yet this result may appear misleading due to the equivalent noise bandwidth (ENBW) of each frequency bin. A more meaningful way to evaluate the measurement is to evaluate the effective bit resolution of the output signal after decimation. This type of evaluation can be simplified if it is assumed that the output is low pass filtered by a brick wall filter with effective bandwidth BW_{wn} , that cuts off at the modulation rate of 140Kb/s (i.e. 140KHz) with the passband noise being white.

For the discriminator the full scale carrier deviation is $\pm 5\text{MHz}$ (10MHz). Also, the full scale modulation deviation is $\pm 50\text{KHz}$ (100KHz). The ratio of these full scale

Figure 4.12 : Measured timing diagram for the $\Delta\Sigma$ frequency discriminator.

deviations is related to the magnitude of the signal power level by

$$S_{dB} = -20 \log \left(\frac{D_{dis,fs}}{D_{mod,fs}} \right) = -40dB \quad (4.28)$$

This relationship assumes that the relative power level of a full scale deviation is zero dB. From Figure 4.13 the relative signal power level S_{dB} in the 70KHz frequency bin is approximately -24dB. This measured result indicates a 12dB scaling difference with respect to the calculated value of equation (4.28). This results from the assumption that a full scale deviation results in a relative power level of zero dB. From the same curve of Figure 4.13, assuming a white noise passband BW_{wn} of 140kHz, the approximate noise power level η_{dB} is -52dBc. In order to calculate the SNR for the modulation frequency bin the ENBW must be determined.

Figure 4.13 : The power spectral density of the captured d.s.criminator output bitstream.

The ENBW per bin can be determined using [Har78]

$$ENBW = \frac{\sum_n w^2(nT)}{\left[\sum_n w(nT) \right]^2} \times \text{binwidth} \quad (4.29)$$

where $w(t)$, the temporal weightings [Nut91], are given by

$$w(nt) = \frac{1}{L} \sum_{n=0}^N a_n \cos\left(\frac{2\pi nt}{L}\right); (|t| \leq L/2) \quad (4.30)$$

and L is the duration of the temporal weightings. Harris [Har78] conveniently computes the coefficient of (4.29), normalized to the bin width, for a minimum 4-sample Blackman-Harris window, to be 2.0. The ENBW now becomes

$$ENBW = 2 \times \frac{f_s}{N} \quad (4.31)$$

where frequency-bin width f_s/N is a function of N , the number of points in the FFT, and f_s is the sampling frequency. From (4.31) ENBW is determined to be 1250Hz. From the measured values and assuming a white noise bandwidth BW_{wn} up to and including the modulation frequency, the signal to noise ratio becomes

$$SNR_{dB} = \frac{10 \log \left[(v_s^2) \times k \right]}{10 \log \left[\left(\frac{v_n^2}{ENBW} \times BW_{wn} \right) \times k \right]} \quad (4.32)$$

$$SNR_{dB} = \frac{S_{dB} + 10 \log(k)}{\eta_{dB} - 10 \log(ENBW) + 10 \log(BW_{wn}) + 10 \log(k)} = 7.51 dB \quad (4.33)$$

where k is a magnitude scaling factor introduced by the software used to evaluate the FFT. This term cancels out in equation (4.33). In order to determine the effective number of bits from 7.51 dB, from Proakis [Pro89] for a pulse code modulated signal, the rms error can be determined using

$$error_{rms, dB} = 20 \log \left(\frac{\Delta}{\sqrt{12}} \right) = 20 \log \left(\frac{2^{-N}}{\sqrt{12}} \right) \quad (4.34)$$

Here Δ is the quantizer step size. Separating and solving for the right-hand side of (4.33) we have

$$error_{rms, dB} = -10.8 - 6N_B \quad (4.35)$$

Where N_B is the number of bits used to represent each sample. Equating the right hand side of (4.34) to the noise level (relative to the signal magnitude) of (4.32) leads to an effective number of bits/sample N_B of approximately 0.55.

This poor result indicates several problems with the circuit. The magnitude of the

signal power level is directly related to the magnitude of the frequency deviation of the modulation signal. Increasing the deviation would lead to greater SNR. This particular discriminator design would also provide better quality discrimination for lower data rates. This problem can be seen in the result of Figure 4.13 where the existence of only a first order quantization noise shaping at the modulation frequency results in high levels of noise and therefore reduced SNR. In the range 500KHz to 1.5MHz the quantization noise has second order shaping, yet below this point frequency discriminator operation degrades to first order operation. This quantization noise degradation dramatically reduces the effective value of N_B over the baseband range. The degradation may be a result of leakage on the integrating capacitors of the charge pump circuit. Under ideal operation each packet of current from I_1 and I_2 causes a voltage step on the integrating capacitors for each cycle of the reference frequency. If these capacitors are loaded by a finite resistance during the reference cycle when the integrating capacitors are neither charging or discharging (i.e. U, D high and $\bar{U}, \bar{D}$ low) then the current can leak off the capacitors with a time constant τ_L given by

$$\tau_L = R_{LOAD} \times C \quad (4.36)$$

This time constant is inversely proportional to the frequency where the chargepump breaks down and ceases to act as an integrator.

These results indicate that this particular implementation of the frequency discriminator is well suited to large deviation, low data rate digital frequency modulation. Considerably better performance is expected from VLSI implementation work being done by Beards [BC92].

Chapter 5

Experimental Investigation for Validation of Broadband FM Conversion

In chapter 2, an experimental testbed was presented (Figure 2.6) for use in testing the concept of single conversion FM broadband mixing. For the fractional-N PLL synthesizers used in the testbed, in static operation, one expects synthesizer frequency outputs with zero frequency and phase error, resulting in a single fixed IF tone after the frequency conversion process. For non-steady-state frequency operation it can be expected that some level of frequency and residual phase error will occur which is a function of the open loop gain of the synthesizer. From linear theory, frequency independent finite open loop gain limitations result in a fixed percentage frequency error of the magnitude frequency deviation of the modulating signal. Frequency dependent finite open loop gain limitations (ie; a function of the PLL phase error transfer function) result in a frequency error which is dependant upon the modulation rate of the modulating signal. In the case of the testbed the magnitude deviation and repetition rate are varied with respect to a wideband sinusoidal FM test vector. Minor frequency differences between the T_x and R_x synthesizers, a function of their respective open loop gains, would directly result in IF distortion.

In this chapter, because ideal synchronization has been assumed, the mechanisms which potentially impair broadband operation and lead to IF distortion are isolated and evaluated with the testbed, using a single Fourier component, a broadband sinusoidal FM test vector. After experimental variables and measurements are defined with respect to the experimental testbed, test results are presented as they relate to key issues pertinent to the operation of the $\Delta\Sigma$ frequency synthesizer and $\Delta\Sigma$ frequency discriminator, as part of a broadband digital FM transceiver.

5.1 Experimental Variables Defined

Using the test set-up of Figure 5.1 (reprinted here from Figure 2.6) the R_x synthesizer was mixed to an IF frequency of 105 MHz. At the T_x end, an offset frequency of 105 MHz was added to the T_x synthesizer which allowed easy evaluation of broadband conversion at an IF center frequency that suited the input frequency range of the $\Delta\Sigma$ frequency discriminator, as indicated by Figure 4.10. At the same time this allowed nominally identical synthesizers to be used at both the T_x and R_x ends. A broadband sinusoidal FM test vector (see section 2.2.3), with no data present, was used as a vehicle to drive both synthesizers, thereby providing a well defined environment for determining the distortion and SNR potential of the broadband conversion architecture. Issues and details regarding the mixers are discussed in Section 5.6.

For the purposes of testing broadband conversion both the T_x and R_x synthesizers of Figure 5.1 were fine synchronized by applying the same 10 MHz reference frequency to both components. The dependant variable measured using the testbed was the IF output signal, designed to be at 105 MHz. The level of time dependent frequency movement with respect to the desired IF center tone, a result of the test vector repetition rate and frequency deviation, was specifically used as a measure of impairment (ie; the analog mismatch between the T_x and R_x synthesizers for

Figure 5.1 : Block diagram of the experimental testbed for broadband coded FM.

various values of the independent variables of the test vector). The independent variables used as controls in the experimental measurements were the repetition rate (no data present), and the peak deviation of the sinusoidal wideband FM test vector, as controlled by the content of the double precision digitally encoded $\Delta\Sigma$ modulated bitstream. In particular, frequency deviations of ± 2.5 MHz, ± 625 KHz, ± 156 KHz, and repetition rates of 1 KHz and 10 KHz were used. Note that in the absence of data being added to the $\Delta\Sigma$ modulator input, since the same sinusoidal test vector is used in the T_x and R_x modulators, the resulting IF signal would be a pure 105 MHz sine wave in the ideal case. The characteristics of the sinusoidal FM test vector are shown in Figure 2.1 and described in section 2.2.3.

5.2 Distortion Analysis for Broadband FM Conversion

Figure 5.2 shows two modified block diagram models of the PLL in terms of its phase error output and DMD phase components. The open loop gain H_{OL} from Figure 5.2 (b) is given by [RC92c]

$$H_{OL} = \frac{k_\theta F(s) k_v}{s} \quad (5.1)$$

The full definition of H_{OL} from equation (3.14), for the PLL synthesizers used, is given by

$$H_{OL}(j\omega) = -\frac{\omega_n^2 k_v k_p}{(n + \beta) \tau_1 \omega^2} \left[\frac{1 + j\omega \tau_2}{j\omega [2\zeta \omega_n + \tau_3 (\omega_n^2 - \omega^2)] + \omega_n^2 - \omega^2 - 2\zeta \tau_3 \omega_n \omega^2} \right] \quad (5.2)$$

The maximum phase error $\theta_{e,m}$ as a function of the maximum phase variation $\theta_{b,m}$ resulting from the broadband sinusoidal FM signal modulating the $\Delta\Sigma$ modulator bitstream $b(t)$ is given by

$$\theta_{e,m} = \frac{1}{1 + H_{OL}} \theta_{b,m} = \frac{1}{1 + H_{OL}} \beta_{FM} \quad (5.3)$$

where β_{FM} was defined in section 2.1.

(a)

(b)

Figure 5.2 : Block diagram of the model of a Fractional-N PLL in terms of its phase components.

This phase error in addition to the error due to path delay will be discussed later (see section 5.4).

The phase variation introduced by the sinusoidal FM test vector is

$$\theta_b(t) = \int_0^t \omega(t) dt = \int_0^t 2\pi\Delta f \sin(2\pi f_m t) dt + 2\pi f_c t \quad (5.4)$$

Performing the integration and dropping the DC gives

$$\theta_b(t) = \frac{\Delta f}{f_m} \sin(2\pi f_m t) \quad (5.5)$$

Figure 5.3 : Phasor representation for analysis of IF broadband conversion distortion.

If the maximum phase deviation due to the sinusoidal FM test vector is considered, then

$$\theta_{b,m} = \frac{\Delta f}{f_m} = \beta_{FM} \quad (5.6)$$

which is the classical definition of the maximum phase deviation for sinusoidal FM. Equation (5.3) shows that the phase error θ_e only approaches zero as the open loop gain H_{OL} approaches infinity. For all practical purposes the circuit gain is finite, and the high pass nature of the phase error transfer function gives decreasing phase error suppression for increasing frequencies. Using this analysis it can be seen that, for finite gain, residual levels of phase error would directly translate into minor frequency errors in the closed loop output of a T_x synthesizer, with respect to an alternate yet similarly constructed R_x synthesizer. Assuming synchronized, non-steady state operation, then the phase difference (path delay independant) between the T_x and R_x synthesizers is given by

$$\Delta\theta_{e,m} = \left[\frac{1}{1 + H_{OL,Rx}} - \frac{1}{1 + H_{OL,Tx}} \right] \theta_{b,m} \quad (5.7)$$

If one views the broadband sinusoidal FM driven T_x and R_x synthesizer outputs as rotating phasors in frequency, as shown in Figure 5.3, then the minor phase differences $\Delta\theta_{e,m}$ existing between synthesizers results in a phase dependant IF frequency F_{IF} , with magnitude bandwidth Δf_{IF} . Ideally if F_{RX} and F_{TX} have the same phase (ie; $\Delta\theta_{e,m}=0$) then the vector difference resulting from the broadband conversion leads to zero IF distortion (a pure tone as symbolized by a point at the origin of the axes). For finite phase differences the magnitude of F_{IF} is directly related to the phase error $\Delta\theta_{e,m}$ and results in a rotating vector at IF with a given distortion bandwidth. Any desired level of SNR at the IF stage must adjust the frequency deviation of a data bit in order to compensate for the magnitude of this frequency distortion.

Alternatively, the frequency imbalance inherent in this type of PLL synthesizer can also be understood by defining the magnitude and phase of the frequency difference that can occur between a desired and a realized output. If $f_{TX,d}(t)$ is defined as the desired output frequency and $f_{TX,a}(t)$ the actual or realized output frequency, then the difference between these components $f_{TX,e}(t)$ can be defined as the frequency error at the synthesizer output due to the filter action, and is given by

$$f_{TX,e}(t) = f_{TX,d}(t) - f_{TX,a}(t) \quad (5.8)$$

and the peak frequency deviation given by

$$\Delta f_{TX,e} = f_{TX,d}(t) \left| \frac{1}{1 + H_{OL,TX} \angle \theta_{OL,TX}} \right| \quad (5.9)$$

These relationships are shown in Figures 5.4 a) and b) and the phase error $\theta_{TX,e}$ between the desired and actual signals defined by

$$\theta_{TX,e} = \angle \frac{1}{1 + H_{OL,TX} \angle \theta_{OL,TX}} \quad (5.10)$$

Therefore the closed loop transfer function becomes the source of the phase and magnitude shift between $f_{TX,d}(t)$ and $f_{TX,a}(t)$ (note that the phase shift caused by additional path delay is an issue discussed separately). If this analysis is carried further, then the difference between two dependant (ie; same reference frequency)

Figure 5.4 : a) The phase relationship between the desired and actual transmit frequency, b) the magnitude frequency error between the desired and actual transmit frequency.

transfer functions in both phase and magnitude, can be viewed as the difference between the T_x and R_x 'actual' frequencies. This difference is in effect the overall frequency deviation error at the particular modulating frequency. β_r or residual β , is the difference in the peak frequency deviation caused by the filtering action of the T_x and R_x PLL sections, and can be defined using the peak frequency errors of both the T_x and R_x synthesizers.

$$\beta_r = \left| \frac{\Delta f_{TX,e} \angle \theta_{TX,e} - \Delta f_{RX,e} \angle \theta_{RX,e}}{f_m} \right| \quad (5.11)$$

Alternatively, using the definition of peak frequency error provided in equation 5.9, equation 5.11 can be expanded to

$$\beta_r = \frac{\left| f_{TX,d}(t) \right| \frac{1}{|1 + H_{OL,TX}|} \angle \theta_{TX,e} - f_{RX,d}(t) \left| \frac{1}{1 + H_{OL,RX}} \right| \angle \theta_{RX,e}}{f_m} \quad (5.12)$$

For a broadband FM conversion process the desired frequency deviations $f_{TX,d}(t)$ and $f_{RX,d}(t)$ at the T_x and R_x synthesizers respectively will be identical, and therefore equation 5.12 can be simplified to

$$\beta_r = \frac{f_{RX,d}(t)}{f_m} \left| \frac{1 \angle \theta_{TX,e}}{1 + H_{OL,TX}} - \frac{1 \angle \theta_{RX,e}}{1 + H_{OL,RX}} \right| \quad (5.13)$$

Equation 5.13 provides a closed form solution for evaluating the peak phase deviation at the IF stage as a function of each PLL open loop gain, FM deviation, and FM rate. The peak phase deviation at the IF output β_{IF} can be evaluated from equation 5.13 by substituting the peak frequency deviation of the FM test vector for $f_{RX,d}(t)$, and providing correct values for the open loop gain, the phase error, and the FM rate.

The non-steady state operating point at which the above phase difference becomes dominant in broadband mixing is a function of the individual open loop gains, which are directly related to the frequency deviation Δf and the repetition rate f_m of the test vector. This is the subject under investigation in section 5.4. From a testing perspective it becomes important to verify the effects of these minor differences on the IF signal during broadband mixing, and at what values of repetition rate f_m and frequency deviation Δf these effects might reduce the IF SNR to levels that are impractical for a digital FM radio receiver. It would be expected that the worst effects would show up at higher levels of f_m and Δf .

The phasor diagram of Figure 5.3 assumes the magnitude of the voltage signals that develop the T_x and R_x frequencies are constant amplitude over the FM bandwidth. Variations in the magnitude synthesizer output voltages over this bandwidth will lead to AM modulation of the resulting IF waveform. The implications of this form of distortion (AM to PM conversion) through subsequent non-linear action will not be addressed in this thesis.

5.3 Frequency Synthesizer Characterization using a Sinusoidal Broadband FM Test Vector

Figure 5.5 shows the power spectrum of the individual T_x and R_x synthesizers in steady-state frequency operation at 815 MHz. The spectrum of each synthesizer indicates loop bandwidths of about 90 KHz and phase noise power levels less optimum than the single sideband phase noise results of Figure 3.17 (due to empirical tuning of loop gain for wideband operation). The contour of the phase noise shows the interaction of the $\Delta\Sigma$ modulator high pass shaped quantization noise, the VCO phase noise, and the closed loop suppressing action of the PLL. For fixed frequency operation the synthesizers provided high quality single tone outputs, similar to Figure 5.5, over most of the operating bandwidth. Using the testbed of Figure 5.1 each synthesizer was characterized for sinusoidal frequency deviations of ± 2.5 MHz, ± 625 KHz, and ± 156 KHz for both 1 KHz and 10 KHz. Figures 5.6, 5.7 and Figures D.1 through D.4 (Appendix D) show the frequency versus time plots of the individual synthesizers in wideband operation. These plots are sinusoidal as expected, aside from the obvious sampling error for the ± 625 KHz and ± 156 KHz at 10 KHz rate signals, shown in Figures D.1 and D.3. Because a synchronizing signal was not used for the individual frequency versus time measurements the apparent phase offset between T_x and R_x outputs has no significant meaning. The power spectrums of Figures 5.8 and Figures D.5 through D.9 all show responses which indicate well matched T_x and R_x synthesizers. These results were generated independently and compared in non-real time and therefore these characterizations can only provide an indication of correct operation of the test setup components for the various test vectors. They cannot provide sufficient evidence of the matching of the complex combination of time constants and gain factors within each of the T_x and R_x phase locked loops. The necessary evidence of synthesizer match/mismatch conditions must come from a controlled examination of the mixing process involving each synthesizer, first mixed with itself and secondly mixed with each other, for the same set of test vectors. Results from such an examination will provide initial insights into a more specific and pragmatic approach for continued research into broadband conversion.

(a)

(b)

Figure 5.5 : Plots of the power spectrums of the individual a) R_x and b) T_x frequency synthesizers operating in fixed frequency mode.

2 of/de 2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
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Figure 5.6 : Plots of frequency versus time for the R_x and T_x frequency synthesizers using a 10 KHz repetition rate and peak deviation of ± 2.5 MHz.

Figure 5.7 : Plots of frequency versus time for the R_x and T_x frequency synthesizers using a 1 KHz repetition rate and peak deviation of ± 2.5 MHz.

(a)

(b)

Figure 5.8 : Plots of the power spectrum for a) the R_x and b) the T_x frequency synthesizers operating with a peak deviation of ± 2.5 MHz and 10 KHz repetition rate.

5.4 Ideal Broadband Conversion Measurement

As discussed in section 2.2.4, for testing purposes ideal synchronization has been assumed between the T_x and R_x sections of the test set-up using a common reference frequency. In this section broadband conversion is investigated under ideal PLL matching conditions by mixing the effective T_x/R_x synthesizer with itself (self mixing). From equation 5.7 the only differences between the synthesizer signals at the mixer T_x and R_x inputs will be the distortion added by the offset frequency mixer, the 920 MHz filter (FBPF-915/30) [KCOM], and the interconnecting semi-rigid coaxial cables. The 3 inch coaxial lines used in the test setup exhibit phase delays of ~ 0.5 ns, while the 920 MHz filter has ~ 55 ns of delay at the center frequency. Test setup sensitivity to these path impairments will show up as a measurement artifact but should be non-dominant with respect to the broadband measurement results. Figure 5.9 shows how ideal broadband conversion can be measured using the test set-up of Figure 5.1. Evaluation of effectively ideally matched synthesizers, although simple in concept, does provide a baseline for qualifying the level of distortion introduced through the use of mismatched synthesizers for broadband conversion.

Figures 5.10 through 5.12 show the measured spectral response of the IF frequency for the self-mixing process when the T_x/R_x synthesizer is driven by a wideband sinusoidal FM code using repetition rates of 1 KHz and 10 KHz, and frequency deviations of ± 2.5 MHz, ± 625 KHz and ± 156 KHz. Note here the wideband code used for testing is the sinusoidal test vector previously described in Figure 2.1 (a) (with no data added). Ideally a pure tone should result at 105 MHz. What is expected is some level of sideband distortion due to the difference in delays between T_x and R_x paths. The magnitude of the peak frequency deviation Δf_{IF} is measured, and hence the effective maximum phase deviation β_{IF} seen at the IF frequency. This can be used as an empirical measure of the quality of the best case broadband conversion process. Table 5.1 presents the measured results from this ideal case. It should be noted that the results in Table 5.1 are the measured results for the T_x synthesizer used as the T_x/R_x synthesizer. The R_x synthesizer results were virtually identical and therefore Table 5.1 and Figures 5.10 through 5.12 represent the measuring sensitivity of the test set-up for each combination of independent variables. Essentially the test set-up of Figure 5.19 can be viewed as a differential delay discriminator, similar to ones used in FSK demodulators. The difference between the received frequency and a delayed version of it, when

TABLE 5. 1 : Ideal Broadband Conversion Results.

Independent Variables			Dependant Variables		% Error
f_m	$\Delta f_{Tx,Rx}$	$\beta_{Tx,Rx}$	Δf_{IF}	β_{IF}	$\Delta f_{IF}/\Delta f_{Tx,Rx}$
1 KHz	± 2.5 MHz	5000	2 KHz	2	0.04%
	± 625 KHz	1250	< 2 KHz	< 2	$\sim 0.04\%$
	± 156 KHz	312	< 2 KHz	< 2	$\sim 0.04\%$
10 KHz	± 2.5 MHz	500	20 KHz	2	0.4%
	± 625 KHz	125	< 20 KHz	< 2	$\sim 0.4\%$
	± 156 KHz	31.2	< 20 KHz	< 2	$\sim 0.4\%$

compared to a decision threshold level, can be used to determine whether a higher or lower frequency was transmitted. The value of the delay required in such systems is a function of the period of the received frequency. In this set-up the delay is fixed by the path between the T_x synthesizer and the delayed version of itself through the offset frequency generation.

Figure 5.9 : Block diagram of the experimental testbed used for evaluation of ideal broadband conversion.

(a)

(b)

(c)

Figure 5.10 : Plots of the power spectrum of the T_x in self mixing for a frequency deviation of a) ± 2.5 MHz, b) ± 625 KHz, and c) ± 156 KHz at a rate of 10 KHz.

(a)

(b)

Figure 5.11 : Plots of the power spectrum of the T_x in self mixing for a frequency deviation of ± 2.5 MHz and a rate of 1 KHz for wide and narrow frequency spans.

(a)

(b)

Figure 5.12 : Plots of the power spectrum of the T_x in self mixing for a frequency deviation of a) ± 625 KHz and b) ± 156 KHz at a rate of 1 KHz

Variations in the period of the T_x signal are a result of changing the repetition rate of the broadband signal from 1 KHz to 10 KHz.

The results of this broadband measurement clearly indicate that, aside from the effect of delay discrimination, wideband deviations as large as ± 2.5 MHz for a 10 KHz repetition rate result in very little perceivable distortion when self mixed.

5.5 Broadband Conversion Measurement

If two separate synthesizers are now considered for the T_x and R_x components, Table 5.2, summarized from Figures 5.13 and 5.14, shows the results of the broadband conversion measurements, for the given sinusoidal FM peak frequency deviations and code repetition rates (no data present on the test vector).

TABLE 5. 2 : Broadband conversion measurement results.

Independent Variables			Dependant Variables		% Error
f_m	$\Delta f_{Tx,Rx}$	$\beta_{Tx,Rx}$	Δf_{IF}	β_{IF}	$\Delta f_{IF}/\Delta f_{Tx,Rx}$
1 KHz	± 2.5 MHz	5000	30 KHz	30	0.60%
	± 625 KHz	1250	8 KHz	8	0.64%
	± 156 KHz	312	2 KHz	2	0.64%
10 KHz	± 2.5 MHz	500	700 KHz	70	14.0%
	± 625 KHz	125	140 KHz	14	11.2%
	± 156 KHz	31.2	40 KHz	4	12.8%

The most obvious conclusion for the results of Table 5.2, considering the ideal results of Table 5.1, is that there are large mismatches between the PLL gains of the T_x and R_x frequency synthesizers. The ideal broadband conversion process provided evidence that broadband conversion works well for ideally matched synthesizers, regardless of peak deviation or repetition rate. This means that modulation rates approaching the bandwidth of the phase locked loop could be implemented if both the open loop gain H_{OL} and the transient response of the transmitting and receiving synthesizers were well matched. Although this is not recommended, the idea was successfully tested with a ± 2.5 MHz sinusoidal test vector with a 50 KHz repetition rate using the ideal conversion process (self

mixing) of Figure 5.9. The results of this specific test will not be presented in this thesis.

The results of Table 5.2 show that changes in peak deviation $\Delta f_{Tx,Rx}$ result in an IF peak phase deviation error that is roughly a fixed percentage of $\Delta f_{Tx,Rx}$. On the other hand changes in the repetition rate result in an error that is a function of frequency (ie; phase error suppression increases with decreasing frequency). Both of these observations are supported by the theory of linear time-invariant systems [Oga70]. What is crucial about these results is the significant levels of IF distortion that have resulted from mismatches in the PLL open loop gain between the T_x and R_x frequency synthesizers and the delays between the synthesizers in the test setup (the latter being non-dominant). Equation 5.13 provides an approximate theoretical closed form solution which suggests the source of the mismatch between the T_x and R_x synthesizers. If a deviation of ± 156 KHz and a repetition rate of 1 KHz is considered as an example, then from Table 5.2, β_{IF} is measured to be 2. If these two values are now applied to equation 5.13, and $H_{OL,RX}$ is assumed to be 20, $H_{OL,TX}$ as 30, $\theta_{RX,e}$ as -30 degrees, and $\theta_{TX,e}$ as -10 degrees, then the calculated magnitude of the residual phase deviation β_r is found to be 4.99. This is well within a magnitude of the measured value.

A qualitative evaluation of SNR based upon these results indicates that the frequency deviation required to transmit a data bit (over the level of mismatch distortion) in order to make the data bit recoverable increases in direct proportion to the FM repetition rate. It should be noted that the level of SNR determined from the measured results would not include real channel distortion. However, real channel distortion would be non-dominant with respect to the distortion levels that have resulted from synthesizer PLL mismatch, for the particular synthesizers used here.

The disparity between the T_x and R_x synthesizers finds its source in the design of their type 2 fifth order phase locked loops. Initially the loops were designed with static LO operation in mind, and therefore additional filtering was introduced to reduce phase noise levels and reference frequency feedthrough. Yet the fifth order design consists of many independent variables (time constants) that affect the gain and delay response of the loop, as shown in the detailed design of section 3.1 in this thesis. Trying to match the open loop gain in a differential environment with discrete components that all have a 10% tolerance has led to differences in the T_x and R_x loops that appear to difficult to resolve in the present prototypes. It should

(a)

(b)

(c)

Figure 5.13 : Plots of power spectrum for the IF frequency using a 10 KHz repetition rate and peak deviations of a) ± 2.5 MHz, b) ± 625 KHz, and c) ± 156 KHz.

(a)

(b)

(c)

Figure 5.14 : Plots of power spectrum for the IF frequency using a 1 KHz repetition rate and peak deviations of a) ± 2.5 MHz, b) ± 625 KHz, and c) ± 156 KHz.

be noted that integrated circuit implementation of the synthesizer are being designed to have considerably higher bandwidths and hence less susceptibility to the problems encountered in the prototypes. In the prototypes, sensitivity to PCB layout due to parasitics, ground plane design, and magnetic cross-coupling also affected the final results. These types of effects are characteristic of a PCB design but not representative of an IC implementation.

An attempt was made to provide tuneability to the overall loop gain by placing a DC clamping circuit (ie; a voltage divider implemented using a trimpot) on the tuning voltage of the VCO of each synthesizer. Independent scaling of the T_x and R_x loop gains provided a technique for empirically minimizing loop gain mismatches. Yet this brute force approach was incapable of fine tuning the independent time delay gain elements of each loop. The minimizing technique was applied each time a test vector was changed and the use of this gain scaling, through the clamping circuit, provided clear evidence that increasing loop gain had the effect of reducing the T_x and R_x synthesizer mismatch. The limitation of the simple clamping circuit lay in the fact that magnitude of all the delay elements around were only scaled by a fixed percent and not tuned independently. The results of Table 5.2 were measured when the differences between the T_x and R_x synthesizers were minimized using the clamping circuit. Therefore the most immediate work that must be done as a next step in investigating the viability of broadband coded FM for wireless communications is exploration of the capability of precisely matching the gains of a set of T_x and R_x frequency synthesizers that will operate with a similar resolution but with a fixed frequency offset. This will require the paring down of the order of the present PLL synthesizer prototypes to reduce the number of matching variables, and then an investigation of how the altered levels of phase noise will affect the SNR of the receiver and/or the radio design, based upon the allocated bandwidth of the wideband channel.

5.6 Qualifying the IF Signal for $\Delta\Sigma$ Frequency Discrimination

In earlier discussions, transceiver operation was based upon the premise that an IF frequency discriminator was available and could be cascaded to the output of the IF mixer. It was unfeasible at this point to test the overall broadband transceiver structure because of the performance limitations at the present state of the research. However, this section makes an attempt to qualify some of the necessary

Figure 5.15 : The power spectral density of the captured discriminator output bitstream.

interface structure required if the $\Delta\Sigma$ frequency discriminator were to be cascaded to the IF output.

From Chapter 4, measured results on the $\Delta\Sigma$ frequency discriminator indicated that its operation was suited to low data rate, wide frequency deviation signals in order to provide adequate SNR at its output. This finding comes from the use of a printed circuit board (PCB) prototype operating with a 10 MHz sampling rate (f_{ref}) on a sinusoidal FM modulation signal of $f_m = 70$ KHz and $\Delta f = \pm 50$ KHz. Figure 5.15 (reprinted from Figure 4.13) shows the power spectral density of the oversampled bitstream. The magnitude of the 70 KHz modulation tone in the spectrum is directly proportional to the peak deviation Δf of the sinusoidal input. A deviation of ± 50 KHz produced a SNR of only ~ 8 dB. In order to realize increased levels of SNR then the peak deviation Δf must also be increased. The constraint of low data rates on the discriminator input is a function of the order of the $\Delta\Sigma$ modulator and the sampling rate. Figure 5.15 results show only first order noise shaping for a second order design. The impairment in the noise shaping reduced the effective dynamic range of the digital signal. If second order noise shaping had been present greater levels of SNR would have resulted through the reduction of the quantization noise floor.

Figure 5.16 : Power spectral density of a $\Delta\Sigma$ frequency discriminator bitstream output, reported by Beards and Copeland [BC92].

Although the results of Chapter 4 indicate poor performance for the printed circuit board prototype (ie; due to charge pump leakage), there is an example of a working VLSI monolithic implementation with much better performance and wider bandwidths [BC92], as indicated in Figure 5.16. If such performance can be assumed, then a radio receiver based upon the architecture of Figure 2.3 becomes more feasible.

The issue of power levels required to drive the frequency discriminator is one that has not been dealt with yet. From Figure 4.10, the discriminator input is the clock input on the dual modulus divider DMD. Data sheets on the divider used, SP8680 [Ples88], specify peak to peak input voltage levels of between 0.4 to 0.8 volts should be supplied to the DMD input. The SP8680 DMD is a 10K compatible ECL component, therefore interfacing a high speed, low level signal requires the use of amplification, *ac* coupling, and level shifting through a 50 Ω matching circuit.

Obviously a purpose-designed DMD would be required in practice that would function with lower input levels (ie; smaller signal swings). Figure 5.17 shows a block diagram of the broadband conversion set-up along with the signal levels at

Figure 5.20 : Block diagram of the broadband conversion set-up including the I/O signal levels.

the I/O ports. The mixers used in the test setup of Figure 5.1, ZLW-2 [MCI], specified a nominal LO input power of +7 dBm and an RF input power of +1 dBm. The actual power levels used in the test setup, shown in Figure 5.17, deviated from the recommended power levels resulting only in increased levels of conversion loss [MCI]. The level of -22 dBm into 50Ω provides a peak to peak voltage of 0.355mv, which could not drive the DMD clock input. As previously mentioned, the DMD input is an ECL component and consists of either an emitter follower input circuit or emitter coupled pair input circuit. This implies a power level with 0.4 to 0.8 volts peak to peak signal swing at current levels capable of driving the base of a bipolar transistor. If the broadband conversion and $\Delta\Sigma$ frequency discriminator circuits were connected together in a receiver configuration, then an amplifier would have to match and amplify the filter $F(s)_2$ output to provide the correct signal levels to drive the DMD input. Alternatively, active mixers with conversion gain could be used. Tuning the gain would help to compensate for losses in the passive filters and allow the provision for optimal signal swing.

In this thesis the $\Delta\Sigma$ frequency discriminator was not connected to the output of the IF stage of the broadband conversion test setup because the performance of the initial breadboard was not up to the standard that would be required. In addition, the 10 MHz reference clock was not available and could not be derived from the reference clocks in use. The derived clock would have needed a DC phase offset function similar to that present on the HP3322B function generators in order to align the sampling frequency of the $\Delta\Sigma$ frequency discriminator so that it sampled after a valid setup time on the clock input to the DMD. This situation impeded an academic evaluation of the combination of the receiver and $\Delta\Sigma$ frequency discriminator as a radio front end. However, knowing the capability of the receiver

to the IF stage and the SNR of the $\Delta\Sigma$ frequency discriminator provided enough information to qualify them as suitable candidates for a fully integrated (ie; small number of external passive components) digital radio receiver. Research work in progress (N. Filiol) will carry this further.

Chapter 6

Conclusions and Recommendations

6.1 Conclusions and Summary

The growing number of users of digital radio services has led to the demand for new and novel architectures that meet application specific requirements in a monolithic environment in both an efficient and economical manner.

The individual contribution of this thesis was the evolution of the higher level radio design and test concept for the proposal of the broadband coded FM application. Contributions also included support work for the evolution of the two components that were investigated as the building blocks for the transceiver, as proposed in Figure 6.1, with synchronization for the receiver as proposed in Section 2.2.2, Figure 2.5. The components were tested separately and in the broadband FM context using a broadband conversion test set-up. The broadband conversion concept, a continuous frequency varying modulation where Δf approaches an infinitesimally small increment (double precision interpolation) in the limit, was tested under both ideally matched synthesizer PLL open loop gain parameters, and poorly matched gain parameters. The results provided insight into the practical requirements for implementing broadband coded FM as an alternative technique for combating the effects of multipath fading in an indoor wireless environment. The results and testbed components additionally provide a foundation for further systems level research into similar areas of broadband modulation for digital radio.

Initially a printed circuit board prototype of a $\Delta\Sigma$ modulated fractional-N frequency synthesizer was constructed and tested in the 800 MHz frequency range, based upon a 10 MHz reference frequency and a type II fifth order PLL. In fixed frequency operation, phase noise levels of -100 dBc/Hz @ 1KHz were achieved, while in fractional-N division operation results degraded slightly, as expected, to -90 dBc/Hz @ 1KHz. Transient behavior to frequency steps showed settling times

Figure 6.1 : Block diagram of a $\Delta\Sigma$ Digital FM transceiver.

on the order of 30 μ s for a 20 MHz frequency step. Measurement of frequency resolution indicated capabilities of 1 Hz movement by toggling the fifth LSB of the $\Delta\Sigma$ modulator input. Typically this type of synthesizer is suited to low data rates as a transmitter due to the limitations of the PLL bandwidth, and is therefore ideally suited to low data rate wide deviation FM applications. Work on evolving monolithic versions are presently in progress using both CMOS and BiCMOS processes [Th93, RCK93].

A printed circuit board prototype of a $\Delta\Sigma$ frequency discriminator was also constructed and tested for a 105 MHz input center frequency, and a 10 MHz sampling frequency. Results for an analog sinusoidal FM input, using a data rate of 50 KHz and a peak deviation of ± 50 KHz, produced output signal-to-noise levels of approximately 9 dB. This low value was the result of an impairment in the noise shaping function of the discriminator. Only first order noise shaping was achieved in the initial design. The degradation in operation was a result of charge leakage from the integrating capacitors of the charge pump circuit. Although the results of the printed circuit board version were not fully adequate for the application, the concept was proven and provided a level of confidence to the innovators (Beards and Copeland) that the VLSI implementation would meet with success. At the time

of this writing initial results for a VLSI implementation were very encouraging, as seen in Figure 5.25. The circuit optimization and SNR expectations for the baseband signal of such a discriminator are presently under investigation by Beards [Bea94].

The results of Chapter 4 indicated that the prototype frequency discriminator was well suited to low data rate high deviation FM applications. The low data rates place the modulation tone further down into baseband where the levels of the shaped quantization noise are non-dominant. The use of wide deviation FM enhances the SNR of the discriminator output signal through the direct relationship between the RF FM peak deviation and the IF peak signal strength.

The availability of two digitally controlled circuit board frequency synthesizer prototypes, one to act as a transmitter and the other for use as a receiver, afforded the opportunity to reach beyond the simple technique of heterodyne conversion and allowed testing of a novel broadband mixing concept. Testing was accomplished using a double precision digitally encoded sinusoidal FM test vector. The IF signal characteristics were measured as a function of the sinusoidal FM repetition rate and peak frequency deviation (no data present on the test vector). The test vectors were varied over a frequency range similar to one that would result from the use of a true pseudorandom broadband code. The results from the experimental testbed evaluation showed that for ideally matched frequency synthesizers, less than 1% error occurred at IF frequencies (an artifact of delay discrimination) regardless of the magnitude frequency deviation. For unmatched frequency synthesizers operating with wide deviation high rate FM signals, the level of mismatch in phase error suppression (ie. a function of open loop gain) led to as much as 15% error at IF frequencies for 10 KHz repetition rates. For practical application this would require that data deviations on the pseudorandom code be greater than 50% of the total error at the IF band, in order to provide the minimum margin necessary in order to detect the presence of a one or zero data bit. Such data deviations would be impossible for present day digital wireless standards (ie; CT2+) due to the bandwidth restrictions (ie; transient behavior) of the synthesizer PLL.

In the test environment, variations in the repetition rate were kept well within the loop bandwidth of the PLL to ensure that frequency mismatch between synthesizers was not a result of differences in the overshoot and settling time of the PLLs. For practical purposes this limited the data rates in the prototype

synthesizers to about 40 KHz, due the 20-25 μ sec settling time of the loop. For more practical data rates the loop bandwidths would have to be opened wider to provide faster settling times. Testing is presently under way in further thesis work (N. Filiol) to properly characterize the broadband concept with suitable precision pseudorandom noise PN code sequence lengths, data rates and deviations to provide optimum SNR at the IF stage [Fi94]. Design of the PN codes based upon the coherence bandwidth for the frequency selective fading indoor wireless environment, and chosen for minimum cross correlation (ie; Gold codes) will lead to greater levels of SNR by combatting the effects of multi-path fading.

If the transceiver block diagram of Figure 6.1 is considered, then experimental results indicate that matching of the T_x and R_x synthesizer open loop gains H_{OL} and phase and frequency response characteristics are fundamental for a broadband FM implementation. It may be possible to use a conventional zero crossing discriminator in the transceiver design, but the $\Delta\Sigma$ frequency synthesizer and $\Delta\Sigma$ frequency discriminator are duals (ie; one generates frequencies from $\Delta\Sigma$ bitstreams, and the other generates $\Delta\Sigma$ bitstreams from frequencies) and therefore complement one another. Both components are suitable for VLSI integration and are easily manipulated to run from the same reference frequency. When combined as shown in Figure 6.1 they are well matched for a wide deviation, low data rate broadband FM transceiver. Yet a transceiver based upon broadband conversion and the concepts presented in this thesis would not be well suited to applications where the T_x and R_x synthesizers were constructed with independant design specifications. The synthesizers must be well matched in both frequency and time domain responses. As a result the concept is better suited to cordless phone applications, especially for multi-channel indoor wireless PBX and conferencing switches. The use of digitally controlled broadband techniques, as outlined in this thesis, alleviates the requirement for a carrier recovery system, and provides a method for a less complex more fully integrated digital radio transceiver.

6.2 Recommendations for Future work on the $\Delta\Sigma$ Fractional-N PLL Frequency Synthesizer

The results for the printed circuit board version of the $\Delta\Sigma$ PLL frequency synthesizer were very promising. Yet the component count was very large, and therefore not very appealing as a printed circuit board product as distinct from a monolithic implementation. The onerous architecture could be reduced in several

ways. First the DMD and MEC circuit could be combined resulting in a single component multi-modulus divider. This would allow a reduction in the gate count of the $\Delta\Sigma$ modulator by removing the concentrator section of the modulator. The complete modulator consisted of two separate third order $\Delta\Sigma$ structures. The Lee and Sodini section [CNL90] was used to concentrate a four bit word to a single bit bitstream output. If the concentrator was removed from the circuit and the DMD designed as a multi-modulus divider, then the entire synthesizer architecture could be reduced by about 20 percent. The cost/benefit strategy of such a design would be more appealing for a VLSI implementation.

For non-static operation of the synthesizer, as in a broadband mixing application, the order of the PLL in the synthesizer has to be reduced. The requirement for well matched synthesizers limits the number of variables that can be introduced into the loop filter because the resulting complex filter response is very difficult to reproduce. Therefore for further broadband testing the loop order must be reduced from a fifth to a second order active loop filter. This will result in higher levels of reference frequency feedthrough and higher levels of double sideband phase noise. The phase noise issue will be fundamental only if the levels are significant enough to impair low end dynamic range and sensitivity, thereby impeding the reception of highly attenuated/faded signals. The reference frequency feedthrough could be ignored because future designs could be based upon higher reference clocks (ie; such as in high speed VLSI implementations) which would experience greater suppression for the same loop filter order and bandwidth.

Research by Bax [Bax94] has led to work in the development of an arithmetically locked loop (ALL) as a spinoff from the $\Delta\Sigma$ PLL frequency synthesizer. The loop can provide digital calculation and control of loop parameters, which would allow a better solution for the matching problems raised in this thesis. The architecture results in a synthesizer with greater potential for monolithic implementation.

Work by Kwasniewski and Thamsirianunt [Th93] has provided promising results towards the development of a CMOS PLL frequency synthesizer (excluding loop filter) that is suitable for GHz range operation. This provides the first step in an initiative towards an all CMOS fully programmable radio transceiver.

One other area that still needs to be investigated for a monolithic fractional-N PLL synthesizer is a fully integrated CMOS $\Delta\Sigma$ modulator. The present version, implemented in an FPGA, suffers with degraded states for DC inputs and only provides a functional bitstream density for a range of 0.25 through 0.75. The

present version needs to have the concentrator circuit removed, and further work done on the MASH3 structure to provide a $\Delta\Sigma$ modulator with higher order noise shaping. It is possible that dithering concepts could be introduced into the architecture to overcome the DC input impairment. This would make a $\Delta\Sigma$ based PLL frequency synthesizer more suitable as a fixed frequency local oscillator in a radio receiver.

6.3 Recommendations for Future work on the $\Delta\Sigma$ Frequency Discriminator

The discriminator design implemented in this thesis was not well suited to printed circuit board techniques. The magnitude of the parasitic capacitive elements associated with transmission lines and soldering pads were well within the magnitude of the charge pump integrating capacitors, making it difficult to use the integrating capacitors on the charge pump as effective tuning elements.

The charge pump leakage current was also a problem that had no apparent solution in the printed circuit board environment. The 2N3904 surface mount transistor arrays suffered from Early voltage problems (V_A) which resulted in a transistor output conductance that allowed current leakage from the integrating capacitors at a rate that affected the second order statistics of the discriminator. More work needs to be done in developing an architecture that is not so dependent on circuit problems in the analog integrator. Presently work has begun on such an architecture and has proceeded past a first iteration in the design cycle [Bax94].

A monolithic implementation of the discriminator by Beards [BC92], based upon his novel design, has provided a clear indication that the technique can be used successfully as a demodulation component in a radio receiver. What still needs to be done is a SNR analysis to show whether this technique, although well suited to VLSI implementation, will provide practical levels of SNR with respect to circuit techniques presently in use.

6.4 Recommendations for Future work on a Digital FM Transceiver for Broadband Applications

As indicated by the results of Chapter 5, more work needs to be done in characterizing the broadband concept with a suitably long enough code sequence

length to determine an analytic solution for designing a broadband transceiver for specified modulation rates and bandwidths.

The block diagram of Figure 2.5 presented a proposal for a broadband coded FM receiver. There is still a large amount of work to be done in implementing the clock recovery and synchronization network required to make the architecture practical. Filiol [Fi94] has begun work in this area.

Other areas discussed but not addressed in this thesis were the issues of noise figure, linearity, compression, dynamic range, and bandlimiting of the RF front end components (ie; the LNA, mixer, and image filter). Although not fundamental to the work in this thesis, they are essential to any practical monolithic digital radio implementation. Analytical issues and novel BiCMOS integrated designs for both an LNA and double balanced mixer are presently under investigation by Long and Copeland [Lon94] and may provide solutions for an integrated analog front end. Image reject filtering, based upon the use of planar on-chip inductors and active tuned circuits, is being investigated by Plett and Khoury [Kho94] in an attempt to provide a monolithic solution to filtering for a fully integrated digital radio.

An issue that has not been dealt with in this thesis, and is mentioned here only because of its importance in radio transceiver design, is the use of waveform coding/decoding and equalization to enhance the level of gain over channel impairments and to improve spectral efficiency. For a full radio transceiver design to be practical, some form of coding or error correction must be provided. For voice transmission, imperceptible levels of noise only reduce voice quality in a gradual manner, yet for data transmission (ie; wireless fax or modem) error rates must be minimized even at the cost of bandwidth (or increased redundancy). The use of amplitude equalization in a broadband receiver may be a requirement to allow constant envelope detection over the entire modulation bandwidth, thereby minimizing AM effects on the IF signal.

With successful work in progress on CMOS and BiCMOS frequency synthesizers, and a successfully demonstrated implementation of a monolithic $\Delta\Sigma$ frequency discriminator, the next step in research is a printed circuit board or multi-chip module MCM radio transceiver implementation which brings the above research ideas together in an attempt to qualify a novel monolithic or hybrid broadband FM transceiver.

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Appendix A

PLL Circuits

This section contains a three page wiring diagram and two page parts list for the PLL presented in Chapter 3.

800 MHz - 1.6 GHz PLL Parts List January 12, 1999

Item	Quantity	Reference	Part
1	11	C1,C2,C3,C4,C6,C7,C8,C28, C29,C49,C50	1.0nf
2	18	C5,C9,C10,C11,C12,C13, C14,C17,C30,C31,C34,C36, C41,C42,C45,C46,C48,C54	0.1μf
3	9	C15,C16,C27,C32,C33,C43, C47,C52,C53	20pf
4	8	C18,C19,C20,C21,C22,C23, C24,C25	47pf
5	2	C26,C38	18nf
6	1	C35	2.2nf
7	1	C37	10μf
8	1	C39	15nf
9	1	C40	100pf
10	1	C44	3.3pf
11	1	C51	27pf
12	3	D1,D2,D3	1N4148
13	1	D4	1N4728

14	20	F1,F2,F3,F4,F5,F6,F7,F8, F9,F10,F11,F12,F13,F14, F15,F16,F17,F18,F19,F20	EMI FILTER
15	1	JP1	H2565
16	4	J1,J2,J4,J5	SMA
17	1	J3	BNC
18	4	L1,L2,L3,L4	1 μ h
19	6	POT1,POT2,POT3,POT4,POT5, POT6	1k Ω
20	11	R25,R26,R27,R28,R29,R30 R31,R32,R34,R35,R36	1k Ω
21	1	Q1	2N3904
22	10	R1,R2,R3,R4,R5,R13,R14, R15,R38,R45	24 Ω
23	9	R6,R7,R9,R11,R17,R20,R21, R24,R41	220 Ω
24	6	R8,R10,R18,R19,R22,R23	100 Ω
25	4	R12,R33,R46,R49	50 Ω
26	1	R16	3k Ω
27	2	R37,R44	10k Ω
28	2	R39,R40	68 Ω
29	2	R42,R43	3.9k Ω
30	2	R48,R47	500 Ω
31	1	T1	T16 - 1
32	2	U1	MACH110
33	1	U2	MC10H124

34	1	U3	CXB112Q
35	1	U4	SP8782
36	4	Q2,Q3,Q4,Q5	MAR 3
37	1	U6	AD829
38	1	U7	UPB581C
39	2	U5,U8	MSC-2-11
40	1	U9	C-616 (VCO)

Appendix B

The Third Order Digital $\Delta\Sigma$ Modulator Circuit

This section contains a wiring diagram and parts list for the printed circuit board design of the digital $\Delta\Sigma$ modulator. It also contains the full set of schematic diagrams for the third order digital $\Delta\Sigma$ modulator design, implemented in a single field programmable gate array (LCA).

LCA Board**January 3, 1999**

Item	Quantity	Reference	Part
1	2	C1,C3	0.1 μ f
2	1	C2	1.0nf
3	3	JP1,JP3,JP4	H2558
4	1	JP2	H2557
5	1	JP5	H3575
6	7	R1,R2,R3,R4,R5,R6,R7	10k Ω
7	4	SIP1,SIP2,SIP3,SIP4	10k Ω
8	1	U1	27512 (Eprom)
9	1	U2	LCA3090 (FPGA)
10	1	U3	MACH110 (PLD)

CC SCH
CARRY CONTROL CIRCUIT

FIG. 5

LINEAR EQUATION IMPLEMENTATION

FIG. 6

Appendix C

$\Delta\Sigma$ Frequency Discriminator Charge Pump Circuits

This section contains a wiring diagram and parts list for both the $\Delta\Sigma$ frequency discriminator and the sub-component charge pump circuit. An additional schematic diagram of the charge pump is provided.

Frequency Discriminator Parts List April 19,1993

Item	Quantity	Reference	Part
1	2	C1,C2	0.1 μ f
2	19	R1,R2,R3,R4,R5,R7,R8,R9, R10,R11,R12,R13,R14,R15, R16,R17,R18,R19,R20	220 Ω
3	1	R6	560 Ω
4	1	U1	SP8680
5	1	U2	CXB1112Q
6	1	U3	Charge Pump
7	1	U4	MC10E1651
8	1	U5	PAL1016RD8

Charge Pump Parts List

July 29, 1993

Item	Quantity	Reference	Part
1	2	C1,C2 1	15pf
2	1	R1	200 Ω
3	4	R2,R34,R36,R38	18 Ω
4	2	R3,R40	390 Ω
5	4	R4,R9,R12,R29	10 Ω
6	3	R5,R10,R13	120 Ω
7	3	R6,R11,R14	820 Ω
8	3	R7,R25,R27	47 Ω
9	1	R8	2.2K Ω
10	1	R15	27K Ω
11	1	R16	1.5K Ω
12	1	R17	100 Ω
13	6	R18,R19,R20,R21,R24,k26	68 Ω
14	1	R22	470 Ω
15	1	R23	150k Ω
16	1	R28	8.2k Ω
17	4	R30,R31,R32,R33	2.7k Ω
18	3	R35,R37,R39	56 Ω
19	6	R41,R42,R43,R44,R45,R46	10k Ω
20	9	U1	MMPQ3904

Charge Pump Schematic
File: PROJECT1\ChargePump
Sheet: 1 of 1
Rev: 1.0

Appendix D

Frequency Synthesizer Characterization Plots based upon a Broadband Sinusoidal FM Test Vector

This section contains both frequency versus time and power spectral density plots of the individual T_x and R_x synthesizers operating using a broadband sinusoidal FM test vector, for several variations on FM rate and deviation.

Figure D.1 : Plots of frequency versus time for the R_x and T_x frequency synthesizers using a 10 KHz repetition rate and peak deviation of ± 625 KHz.

Figure D.2 : Plots of frequency versus time for the R_x and T_x frequency synthesizers using a 1 KHz repetition rate and peak deviation of ± 625 KHz.

Figure D.3 : Plots of frequency versus time for the R_x and T_x frequency synthesizers using a 10 KHz repetition rate and peak deviation of ± 156 KHz.

Figure D.4 : Plots of frequency versus time for the R_x and T_x frequency synthesizers using a 1 KHz repetition rate and peak deviation of ± 156 KHz.

(a)

(b)

Figure D.5 : Plots of the power spectrum for a) the R_x and b) the T_x frequency synthesizers operating with a peak deviation of ± 2.5 MHz and 1 KHz repetition rate.

(a)

(b)

Figure D.6 : Plots of the power spectrum for a) the R_x and b) the T_x frequency synthesizers operating with a peak deviation of ± 625 KHz and 10 KHz repetition rate.

(a)

(b)

Figure D.7 : Plots of the power spectrum for a) the R_x and b) the T_x frequency synthesizers operating with a peak deviation of ± 625 KHz and 1 KHz repetition rate.

(a)

(b)

Figure D.8 : Plots of the power spectrum for a) the R_x and b) the T_x frequency synthesizers operating with a peak deviation of ± 156 KHz and 10 KHz repetition rate.

(a)

(b)

Figure D.9 : Plots of the power spectrum for a) the R_x and b) the T_x frequency synthesizers operating with a peak deviation of ± 156 KHz and 1 KHz repetition rate.

END

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FIN