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A SAW-Based Commutation Signaling Modem for Broadband Indoor Wireless Communication

George Kipens, B. Eng.
Department of Electrical Engineering
McGill University, Montreal

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

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Abstract

Commutation signaling is a bandwidth expanding modulation scheme that is robust to multipath induced intersymbol interference. Surface Acoustic Wave (SAW) technology is very suitable for implementing high-speed commutation signaling modems. A SAW-based commutation signaling modem for broadband indoor wireless communication is described. The differential coherent detector is a key element since it performs IF matched filtering and differentially coherent demodulation of the received signal. A commutation signaling differential coherent detector has been implemented using SAW and analog silicon technologies. The bit rate is 40 Mb/s, the chip rate is 200 Mc/s and the operating frequency (IF frequency) is 1 GHz. A multi-carrier commutation signaling modem uses frequency-division multiplexed subcarriers to combine a number of lower rate bit streams in order to achieve high-rate data transmission. A computer simulation study has been carried out to evaluate the performance of multicarrier commutation signaling over a nonlinear multipath channel.

Résumé

La signalisation par commutation est un mode de modulation à étalement de bande qui résiste au brouillage intersymbole dû à la propagation par trajets multiples. La technologie des ondes acoustiques de surface (SAW) se prête très bien à l'implantation d'un modem à signalisation par commutation haute vitesse. Cette thèse décrit un modem de signalisation fondé sur cette technologie et destiné aux communications intérieures sans fil à large bande. Le détecteur différentiel cohérent qui est un élément clé de ce modem assure le filtrage adapté IF et la démodulation cohérente différentielle du signal reçu. On a implanté un détecteur cohérent différentiel à signalisation par commutation dont la conception repose sur les technologies des ondes acoustiques de surface et des dispositifs analogiques au silicium. Le modem a un débit binaire de 40 Mb/s et un débit d'éléments de 200 Mc/s; sa fréquence d'exploitation (fréquence IF) est de 1 GHz. Dans un modem à signalisation par commutation à porteuses multiples, la répartition en fréquence de sous-porteuses multiplées permet de combiner plusieurs trains binaires de faible débit pour obtenir un débit de transmission élevé. On a donc procédé à une simulation informatique pour évaluer le rendement de la signalisation par commutation à porteuses multiples dans une voie multiple non linéaire.

Preface

This research was carried out as part of the collective CITR research effort on Broadband Indoor Wireless Communication. This work was supported by a grant from the Canadian Institute for Telecommunications Research (CITR) under the NCE program of the Government of Canada.

This thesis is organized in two parts. Part I describes a SAW-based commutation signaling modem for broadband indoor wireless communication. Part II presents the results of a computer simulation study on the performance of multicarrier commutation signaling over a nonlinear multipath channel. The research for Part I was done in collaboration with Nortel Wireless Networks (formerly part of Bell-Northern Research), and was co-supervised by Prof. Harry Leib of McGill University and Dr. John Saw of Nortel. The research for Part II was done at McGill University and was supervised by Prof. Harry Leib.

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Glossary of Terms

BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CITR	Canadian Institute for Telecommunications Research
CRC	Communications Research Centre
DIP	Dual-in-Line Package
DT	Designer Test
DPSK	Differential Phase Shift Keying
DQPSK	Differential Quadrature Phase Shift Keying
EHF	Extremely High Frequency
EMFT	Electromagnetic Feed-Through
FFT	Fast Fourier Transform
GL	Generalized Likelihood
IC	Integrated Circuit
IDT	Interdigital Transducer
IF	Intermediate Frequency
IMD	Inter-Modulation Distortion
ISI	Intersymbol Interference
MCM	Multi-Chip Module
NCE	Network of Centres of Excellence
RF	Radio Frequency
SAW	Surface Acoustic Wave
SNR	Signal-to-Noise Ratio
TTI	Triple-Transit Interference
VNA	Vector Network Analyzer

Chapter 1

General Introduction

The continuing large-scale deployment of wireless access technologies, rapid advances in personal computing and wide-spread access to the information superhighway are market forces that are stimulating a demand for tetherless and ubiquitous access to a wide range of high-speed data communication services [1], [2]. These market forces are driving a convergence of wireless communications and broadband access technologies [3], [4]. Important classes of broadband services include multimedia applications, high-resolution document retrieval and video teleconferencing. Broadband wireless access will allow low-mobility users to access large bandwidth-on-demand by means of tetherless multimedia terminals.

This research was carried out as part of the CITR major project on Broadband Indoor Wireless Communication. The main objective of this major project is the system architecture design and demonstration of a broadband indoor wireless digital communication system. This system would provide millimeter-wave access to an ATM LAN at transport bit rates up to about 160 Mb/s. The indoor millimeter wave radio environment can be characterized as a fading multipath channel [5], [6]. The measured delay spreads are on the order of 10-120 ns (i.e., up to about 20 bit intervals) [4]. As a consequence, the main impairment to high-speed data transmission over the indoor radio channel is the multipath induced intersymbol interference (ISI).

Commutation signaling is a bandwidth expanding modulation scheme that is robust to multipath induced ISI [7]. Surface Acoustic Wave (SAW) technology is very suitable for implementing high-speed commutation signaling modems [8]. This thesis is organized in two parts. Part I describes a SAW-based commutation signaling modem for broadband indoor wireless communication. Part II presents the results of a computer simulation study

of the performance of multicarrier commutation signaling over a nonlinear multipath channel.

The rationale for dividing this thesis into two parts is as follows. The differential coherent detector is a key element of the modem since it performs IF matched filtering and differentially coherent demodulation of the received signal. The main objective for Part I of this thesis is to demonstrate the practical feasibility of implementing a commutation signaling differential coherent detector using SAW and analog silicon technologies. The performance targets are as follows: (1) a bit rate of 40 Mb/s, (2) a chip rate of 200 Mc/s, and (3) an operating frequency (IF frequency) of 1 GHz. The chip rate is based on a bandwidth spreading factor of 5 (i.e., 5 chips per bit). This bandwidth spreading factor is a good tradeoff between anti-multipath performance and bandwidth efficiency [8]. The chip rate of 200 Mc/s is determined by SAW device fractional bandwidth considerations (20% target). The IF frequency constraint of 1 GHz is imposed by current SAW fabrication limitations (photolithography linewidth limit of 1 micron).

A multicarrier commutation signaling modem uses frequency-division multiplexed subcarriers to combine a number of lower rate bit streams to achieve high-rate data transmission. In order to achieve an aggregate bit rate of 160 Mb/s, it is necessary to use four subcarriers, each supporting a data rate of 40 Mb/s. Whenever multicarrier modulation is used, the effect of intermodulation distortion (IMD) needs to be considered. In most radio systems, the primary source of intermodulation distortion is the power amplifier. In order to reduce IMD to an acceptable level, the power amplifier needs to be backed off. The amount of backoff required determines the cost and complexity of the transmitter. Part II of this thesis presents the results of a computer simulation study of the performance of multicarrier commutation signaling over a nonlinear multipath channel. Results that show the effect of subcarrier spacing and intermodulation distortion on modem performance are presented by means of eye diagrams and error rate curves.

Part I

A SAW-Based Commutation Signaling Modem for Broadband Indoor Wireless Communication

Chapter 2

Introduction - Part I

Part I of this thesis considers an application of SAW technology to broadband indoor wireless communication over the millimeter wave radio channel. Broadband indoor wireless communication can be categorized as providing low-mobility, tetherless, high-speed data communication within a local area (e.g., office building, university, hospital, etc.). Broadband wireless communication necessitates the use of wide transmission bandwidths. Consequently, the millimeter wave band is considered the most appropriate for this application since it is a relatively large, unused portion of the radio spectrum.

The indoor millimeter wave radio environment is characterized by a large number of propagation paths with different delays and attenuation [5], [6]. Therefore, the major obstacle to digital transmission at high data rates over the indoor radio channel is the multipath induced intersymbol interference (ISI). Commutation signaling was introduced by Turin [7] as an effective technique for combating multipath induced ISI. Commutation signaling can be realized as a direct sequence spread-spectrum technique that uses, in a cyclical manner, a number of different commutation code sequences to spread successive message bits. The commutation code set is chosen to be large enough to ensure that the channel response to any particular code sequence has vanished prior to that code sequence being reused. Individual code sequences are chosen to have low partial cross-correlation. As a consequence, any two message bits arriving simultaneously at the receiver (by any path) will be close to orthogonal. Thus, intersymbol interference is reduced.

Commutation signaling also has an important advantage when technology and implementation issues are considered. In [8], an approach was suggested for implementing a commutation signaling modem based on using

surface acoustic wave technology. SAW technology is capable of providing complex IF signal processing functions and is compatible with semiconductor microfabrication techniques.

Part I of this thesis is organized as follows. Chapter 3 provides the background on commutation signaling. Chapter 4 describes a SAW-based commutation signaling modem for broadband indoor wireless communication. Chapter 5 discusses technology and implementation issues. The differential coherent detector is a key element of the modem since it performs IF matched filtering and differentially coherent demodulation of the received signal. A commutation signaling differential coherent detector has been implemented in multi-chip module (MCM) form using SAW and analog silicon integrated circuit technologies. The performance of the differential coherent detector is determined, to a large extent, by the performance of the SAW devices used for matched filtering. The SAW devices were fabricated on 128° YX-lithium niobate using four different metal systems. Chapter 6 presents measurement results for phase-coded SAW filters operating at a 200 Mb/s chip rate and a 1 GHz IF frequency. This chapter also discusses the effect of metallization on device performance and presents measurement results for the commutation signaling differential coherent detector. Chapter 7 summarizes the work done for Part I.

Chapter 3

Commutation Signaling for Broadband Indoor Wireless Communication

The indoor millimeter wave radio environment can be characterized as a fading multipath channel [5], [6]. The measured delay spread Δ is typically in the range 10-120 ns (i.e., 2-20 bit intervals), when considering a transport bit rate of about 155 Mb/s (SONET OC-3). As a consequence, the main impairment to high-speed data transmission over the indoor radio channel is the multipath induced ISI.

Three known techniques for combating intersymbol interference are adaptive equalization, antenna sectorization and anti-multipath modulation. Adaptive equalization at high data rates is computationally intensive and leads to a complex receiver implementation. Antenna sectorization is an efficient anti-multipath technique. There is, however, a tradeoff between system complexity and the number of antenna sectors needed to combat multipath while maintaining good coverage. Commutation signaling is an effective anti-multipath modulation technique [7]. It was proposed [8] that a combination of antenna sectorization and commutation signaling will provide adequate anti-multipath performance without incurring a substantial penalty in system complexity.

In order to understand how commutation signaling reduces ISI, consider first the transmission of a single spread-spectrum pulse over both a direct and a multipath channel. Fig. 3.1(a) shows the response of a filter matched to a single spread-spectrum pulse that has been transmitted over a direct path. It is assumed that the spread-spectrum signal consists of a number of chips, each of duration T_c . Let the transmitted signal be

$$s(t) = u(t), \quad 0 \leq t \leq T \quad (3.1)$$

where T is the symbol period. Since the filter is matched to the input signal, it

Figure 3.1. Matched filter responses. (a) Single spread-spectrum pulse transmission over a direct path. (b) Single spread-spectrum pulse transmission over a multipath channel. (c) Transmission of two symbols over a multipath channel with $\Delta < T$. (d) Transmission of two symbols over a multipath channel with $\Delta > T$.

has the following impulse response

$$h(t) = u(T - t) \quad (3.2)$$

The output of the matched filter is given by the convolution integral

$$y(t) = u(t) * h(t) = \int_{-\infty}^{\infty} u(\tau) h(t - \tau) d\tau = \int_{-\infty}^{\infty} u(\tau) u(T - t + \tau) d\tau = R_u(t - T) \quad (3.3)$$

where

$$R_u(t) = \int_{-\infty}^{\infty} u(\tau + t) u(\tau) d\tau \quad (3.4)$$

is the autocorrelation function of $u(t)$. Therefore, the matched filter output $y(t)$ corresponds to a time shifted (by symbol period T) version of the autocorrelation function $R_u(t)$. Since $u(t)$ is a spread-spectrum waveform, its autocorrelation function is characterized by a narrow mainlobe of duration T_c , surrounded by lower level sidelobes. The peak of the autocorrelation function occurs at time $t = T$.

Fig. 3.1(b) shows the matched filter response when a single spread-spectrum pulse is transmitted over a multipath channel with three delay paths. The multipath delay spread is Δ . The channel can be modeled as having the impulse response

$$h_c(t) = \sum_{i=0}^2 \alpha_i \delta(t - \Delta_i) \quad (3.5)$$

where α_i is the attenuation factor for the signal received via the i th path and Δ_i is the propagation delay for the i th path. The output of the matched filter is then given by

$$\begin{aligned} y(t) &= u(t) * h_c(t) * u(T - t) \\ &= u(t) * u(T - t) * h_c(t) \\ &= R_u(t - T) * h_c(t) \\ &= \sum_{i=0}^2 \alpha_i R_u(t - T - \Delta_i) \end{aligned} \quad (3.6)$$

The matched filter response now contains three autocorrelation mainlobes,

one for each path. The relative amplitude of each mainlobe is proportional to the strength of each delay path. The autocorrelation mainlobes are spread over the time interval $[T, T + \Delta]$ of duration Δ .

Now, consider the response of the matched filter when two spread-spectrum pulses, corresponding to data symbols a_0, a_1 , are transmitted in succession. Fig. 3.1(c) shows the case when the delay spread Δ is smaller than the symbol period T (i.e., $\Delta < T$). For this case, peak interlacing does not occur. The multipath peaks corresponding to each transmitted symbol can be easily combined by integrating the matched filter response over a time interval of duration Δ . Hence, the transmitted symbols can be detected with little or no intersymbol interference. When peak interlacing is not present, the channel multipath delays can be estimated. Therefore, RAKE multipath combining can be employed [9]. Fig. 3.1(d) shows the opposite condition (i.e., $\Delta > T$). For this case, interlacing of multipath peaks does occur, resulting in significant ISI. This is due to the fact that the multipath channel is still “ringing” from previously transmitted symbol(s) during the multipath combining interval for the current signal.

Commutation signaling is an anti-multipath modulation scheme that expands bandwidth. The principle of commutation signaling is to avoid multipath peak interlacing by choosing a large enough signal set to ensure that the channel response to any particular transmitted signal has vanished prior to that signal being retransmitted. As a consequence, the multipath peaks associated with successive transmissions of the same commutation signal do not overlap. By using multipath diversity combining, commutation signaling exploits the time diversity that is inherent in a multipath propagation medium. The anti-multipath performance of commutation signaling depends on the channel multipath delay spread, the bandwidth expansion factor and the type of multipath combiner employed (e.g., RAKE, post detection integration, etc.).

Fig. 3.2 is a conceptual illustration of a front-end stage for a commutation signaling demodulator. The front-end stage consists of a bank of matched filters. Each filter is matched to one commutation signal. For illustrative purposes, it is assumed that the commutation signal set contains four signals $\{u_0(t), u_1(t), u_2(t), u_3(t)\}$, with vanishing cross-correlation. Each signal has the

same number of chips, of duration T_c . The transmitted data sequence is $a_0, a_1, a_2, \dots$, with $a_i = \pm 1$. It is also assumed that the multipath channel contains three delay paths, as shown in Fig. 3.2. It is seen that each transmitted commutation signal gives rise to three distinct multipath peaks (mainlobes) in the response of its associated filter. The peaks are spread over the time interval Δ . The width of these mainlobes is small compared to the symbol period T because of the bandwidth spreading nature of the commutation waveforms. The same transmitted signal simultaneously causes cross-correlation noise in the responses of the other matched filters. Since the signals $u_i(t)$ have minimal partial cross-correlation, the cross-correlation noise consists of low-level sidelobes only. Referring again to Fig. 3.2, it is seen that, even though the multipath delay spread Δ is greater than the symbol period T , the autocorrelation mainlobes at the output of each matched filter are not interlaced. For example, the multipath peaks associated with symbols a_0 and a_4 are not interlaced. Although both a_0 and a_4 were transmitted using commutation signal $u_0(t)$, peak interlacing is avoided since $\Delta < 4T$.

Figure 3.2. Matched filter responses to transmitted commutation signals.

Figure 3.2 illustrates the use of four commutation sequences. In general, the number of sequences and the length of each sequence are determined from system level considerations. Once the bit rate is chosen, the number of sequences is determined by the channel delay spread. The sequence length (or spreading gain) is determined by a tradeoff between available bandwidth and system error rate performance. Shorter sequences have high cross-correlation sidelobes resulting in high self-generated noise. Longer sequences have lower cross-correlation sidelobes (i.e., lower self-generated noise), but require a larger transmission bandwidth. This topic is discussed further below.

An obvious way of increasing the transmission rate over a multipath channel is to use M -ary modulation with a sufficiently large M . In an M -ary modulation scheme, the information source emits a sequence of symbols from an alphabet that consists of M symbols. For M -ary transmission using an information rate of R bits/sec, the multipath delay spread limits the maximum transmission rate to $R < \log_2 M / \Delta$. For a given R and Δ , the size of the signaling alphabet must be

$$M = 2^{\lceil R\Delta \rceil} \quad (3.7)$$

where $\lceil x \rceil$ denotes the smallest integer greater than x . Thus, for M -ary modulation, the size of the signaling alphabet *increases exponentially* with the multipath delay spread Δ .

$M \times N$ commutation signaling uses N M -ary signal sets. Thus, the total number of signals required is MN . The signals are chosen to be shift quasi-orthogonal. Two signals are said to be shift quasi-orthogonal if their cross-correlation is much smaller than their energy for all time shifts. The value of N is chosen to be large enough to ensure that the channel response to any particular signal set has vanished prior to that signal set being reused. Hence, $N = \lceil R\Delta / \log_2 M \rceil$, and the total number of signals required is

$$MN = M \left\lceil \frac{R\Delta}{\log_2 M} \right\rceil \quad (3.8)$$

Therefore, for commutation signaling, the size of the signaling alphabet *increases linearly* with the multipath delay spread Δ . Since there is a direct relationship between alphabet size and implementation complexity, commu-

tation signaling has an important advantage when technology and implementation issues are considered. Furthermore, for commutation signaling, the signal-to-noise ratio (SNR) performance is determined by the signal constellation in each signal set, while the anti-multipath performance is determined separately by the number of signal sets employed. Hence, commutation signaling provides the flexibility to independently optimize each of these performance criteria.

In [10], a design methodology is described for constructing good commutation signal sets when the multipath delay spread is less than a small integer multiple of the symbol duration. Good codes are selected on the basis of the following two criteria: (1) minimal cross-correlation between codewords for all time shifts within the expected multipath delay spread; and (2) an error probability performance degradation of not more than 3dB when compared against ideal two-fold diversity over a Rayleigh fading channel. As an example, the following codebook of size four and length seven can be used with a delay spread of nearly four times the symbol duration:

$$\{1100000, 0010000, 1001010, 01110110\}$$

Since the delay spread Δ usually varies with the size and type of indoor radio environment, the choice of N may not be optimal for all indoor radio environments. There are two possible approaches to this problem: (1) use a fixed value of N (based on a broad statistical average of rms delay spread measured in different types of indoor radio environments) and optimize the antenna sectorization and selection diversity for individual environments; or (2) use a different value of N for different indoor radio environments. The first approach results in a simpler hardware design but more complex antenna system design. The second approach results in a more complex hardware design but simpler antenna system design. Ultimately, the best approach is the one that provides the most cost-effective and practical solution.

Chapter 4

A SAW-Based Commutation Signaling Modem

Phase-coherent communication at EHF over multipath fading channels is a difficult challenge. In order to alleviate some of the difficulties associated with coherent communication, a commutation signaling scheme that uses binary differential phase shift keying (DPSK) and differential coherent detection was proposed as being suitable for broadband indoor wireless systems [10]. An important advantage of DPSK commutation signaling is that it reduces cross-correlation noise when short spreading sequences are used. Another very suitable scheme uses 4-ary phase sequences and differential detection (i.e., DQPSK). 4-ary phase sequences for commutation signaling are described in [11]. It has been shown that these sequences provide a significant gain in performance when compared with binary sequences. However, the SAW implementation of matched filters is more complex since it requires the use of quadrature SAW devices. The next step in the development of SAW filters for commutation signaling modems could consider this issue.

Fig. 4.1 shows the block diagram for the commutation signaling DPSK modulator. The original data sequence $a(k) = \pm 1$ is differentially encoded into $b(k) = b(k-1) \cdot a(k)$. Thus, phase transitions between successive encoded data bits represent the original data sequence. Each encoded data bit is then modulated by a commutation signaling codeword. The commutation signaling codewords are bandwidth spreading sequences chosen from the following codebook of size three and length five:

$$\{01101, 01111, 10001\}$$

Fig. 4.2 shows the block diagram for the commutation signaling DPSK demodulator. The input to the demodulator is derived from the millimeter wave radio receiver IF stage. The differential coherent detector is a key

Figure 4.1. Commutation signaling DPSK modulator.

Figure 4.2. Commutation signaling DPSK demodulator.

element of the demodulator since it performs IF matched filtering and differentially coherent demodulation of the received signal. Each output from the differential coherent detector is connected to a multipath combiner. The number of output lines corresponds to the number of commutation waveforms. The signal at each output contains narrow autocorrelation mainlobes. The purpose of each multipath combiner is to sum the mainlobes associated with each received symbol in order to obtain diversity gain. Each multipath combiner output is sampled and a decision is made as to whether a binary 0 or 1 was transmitted. The demodulated data from all the multipath combiners is then multiplexed into a serial bit stream.

In [12], the Generalized Likelihood (GL) principle was used to derive the optimal receiver for the broadband indoor radio channel. The GL approach naturally combines channel estimation with data detection and does not require an a-priori channel model. For GL detection, the receiver uses a generalized likelihood ratio test to make a decision as to whether a binary 0 or 1 was transmitted. The decision variables Z_m for a GL receiver that uses spread-spectrum signaling are given in complex envelope form by

$$Z_m = \sum_{i=0}^{I-1} \left| \int_0^{T_o} \bar{r}(t) \bar{s}_m^*(t - \hat{\tau}_i) dt \right|^2 \quad (4.1)$$

where

Z_m denotes the m th decision variable;

I denotes the number of delayed paths in the channel impulse response;

T_o denotes the observation interval;

$\bar{r}(t)$ denotes the complex envelope of the received signal;

$\bar{s}_m(t)$ denotes the complex envelope of the m th wideband transmitted signal;

$\hat{\tau}_i$ denotes the maximum likelihood estimate of the propagation delay for the i th delayed path.

The GL receiver compares the value of Z_m against a fixed threshold and makes a decision in favor of the maximum.

From equation (4.1) it is seen that, with this detection technique, only the multipath delays need to be estimated. Fig. 4.3 presents the structure that implements the decision variable given by (4.1). It consists of a matched filter

Figure 4.3. The Generalized Likelihood (GL) receiver for the broadband indoor radio channel.

followed by a square-law device, a fixed time-delay module T_D , and a RAKE receiver. The purpose of the time-delay module is to allow time for the channel estimator to estimate the channel path delays. Each path produces a corresponding peak in the matched filter output. The channel estimator estimates the path delays by locating the position of these multipath peaks. The path delay estimates $\hat{\tau}_0, \hat{\tau}_1, \dots, \hat{\tau}_{L-1}$ are then used to activate the appropriate taps of the RAKE receiver in order to combine (sum together) the multipath peaks. The output of the equal weight RAKE combiner is sampled at the instant when the multipath peaks are aligned.

The decision variables for a GL receiver that uses commutation signaling and differential decoding can be derived from (4.1) and are given by [12]

$$Z_m = \sum_{i=0}^{L-1} \text{Re} \left\{ \left[\int_0^{T_D} \tilde{r}(t) \tilde{u}_{m-1}^*(t - \hat{\tau}_i) dt \right] \cdot \left[\int_0^{T_D} \tilde{r}(t) \tilde{u}_m^*(t - \hat{\tau}_i - T) dt \right]^* \right\} \quad (4.2)$$

In equation (4.2), $\tilde{u}_{m-1}(t)$ and $\tilde{u}_m(t)$ represent the complex envelopes of the $(m-1)$ th and m th commutation signaling waveforms. Fig. 4.4 shows the

Figure 4.4. GL receiver front-end for differentially encoded binary commutation signaling.

front-end of a GL receiver that forms the decision variable given by (4.2). The front-end consists of a pair of matched filters, blocks that implement delay and complex conjugation, a complex multiplier block and a block that extracts the real part of a complex signal. The GL receiver front-end performs matched filtering and differential coherent detection of the received signal.

Figure 4.5 SAW-based commutation signaling differential coherent detector.

Fig. 4.5 shows the structure of the SAW-based commutation signaling differential coherent detector. This structure uses a passband version of the GL receiver front-end shown in Fig. 4.4. SAW technology is used to implement IF signal processing functions, specifically, correlation and delay. Analog silicon (i.e., RF IC) technology is used to implement the high-speed bilinear multipliers needed for differential coherent detection as well as the low-impedance buffers used to drive the multipliers. The low-impedance buffers help reduce spurious reflections caused by the SAW device metal film interdigital transducers (IDTs).

For differentially encoded binary commutation signaling that uses N commutation waveforms, the differential coherent detector consists of a bank of N SAW matched filters, N SAW delay lines and N bilinear multipliers. Each SAW filter is matched to one of the commutation signals. Differentially coherent detection is achieved by multiplying the output of one matched filter by the delayed (by symbol period T) output of the matched filter corresponding to the previously transmitted symbol. The low pass filters shown explicitly in Fig. 4.5 are sometimes contained within the multipath combiners themselves. The purpose of each low pass filter is to remove the double frequency component after multiplication.

Chapter 5

Technology and Implementation

Phase-coded SAW devices can be used to generate and detect binary phase shift keyed (BPSK) waveforms [13]. Fig. 5.1 shows the structure of a phase-coded SAW device for correlation of a 01101 code. The input IDT is uniform with length vT_c , where v is the SAW velocity and T_c is the chip duration. The output IDT consists of an array of short uniform transducers (taps) which are connected to two bus bars. The separation between output taps corresponds to one chip period sampled by the input IDT. The taps are coded by phase-reversing connections to the bus bars. Fig. 5.1 also shows the response of the output IDT when excited by a rectangular acoustic pulse of one chip duration. This rectangular pulse propagates along the coded transducer and exits each tap in sequence producing the coded waveform. The rectangular acoustic pulse is generated by electrical impulse excitation of the input IDT. The impulse response of the SAW matched filter is a time-reversed replica of the BPSK sequence corresponding to code 01101.

Various SAW IDT structures can be used to implement the matched filters and delay lines shown in Fig. 4.5. One possible scheme for $N=3$ and $M=5$ is illustrated in Fig. 5.2. This embodiment uses a matched filter pair structure consisting of narrowband input IDTs and wideband tapped coded output IDTs. The matched filter bank shown in Fig. 5.2 is comprised of three matched filter pairs. Each matched filter pair has two narrowband input IDTs and two

Figure 5.1. Phase-coded SAW device for correlation of a 01101 code.

Figure 5.2. Phase-coded SAW matched filter bank (wideband tapped output IDTs).

wideband tapped coded output IDTs. Each coded output IDT has 5 taps, corresponding to the number of chips in each commutation codeword. Each matched filter pair ideally provides identical outputs that are delayed in time by one symbol period $T_s = 25$ ns. The separation between taps in each coded output IDT corresponds to one chip period sampled by the input IDT. The taps have a uniform spacing vT_c , where v is the SAW velocity and $T_c = 5$ ns is the chip duration. These taps are identical except that some are inverted in accordance with the required code. The output IDTs in each matched filter pair are separated by distance vT_s , where $T_s = 25$ ns is the symbol time.

Fig. 5.2 presents a *wideband* structure. Another possible embodiment for $N=3$ and $M=5$ is illustrated in Fig. 5.3. This scheme employs a matched filter pair structure consisting of wideband input IDTs and narrowband coded output IDTs. Each matched filter pair has two wideband input IDTs and two narrowband phase-coded output IDTs. The fingers in each output IDT sample

Figure 5.3. Phase-coded SAW matched filter bank (narrowband coded output IDTs).

the impulse response of the corresponding phase-coded waveform. Phase reversals occur at intervals vT_c . The output IDTs in each matched filter pair are separated by distance vT_s . Thus, each matched filter pair ideally provides two identical outputs, of which one is delayed by a symbol period. The embodiment shown in Fig. 5.3 is called a *narrowband* structure.

The phase-coded SAW devices were fabricated on 128°-rotated, Y-cut, X-propagating lithium niobate. This substrate was chosen because of its weak coupling to bulk acoustic waves, its high velocity v and its high electromechanical coupling constant K^2 . A 200 Mc/s chip rate at a 1 GHz IF frequency corresponds to a fractional bandwidth of 20% (i.e., 5 carrier cycles per chip). The maximum intrinsic fractional bandwidth $(\Delta f / f_0)_I$ of a SAW transducer is defined as [13]

$$(\Delta f / f_0)_I = \sqrt{\frac{4K^2}{\pi}} \quad (5.1)$$

This is the widest bandwidth for which it is theoretically possible to keep the insertion loss of the transducer at 6 dB. In order to maximize $(\Delta f / f_0)_I$, K^2 should be as large as possible. 128° YX-lithium niobate is the preferred substrate material for wide bandwidth filters since it has a large value of K^2 (5.3%). For 128° YX-lithium niobate, $(\Delta f / f_0)_I$ is 26%.

The acoustic wavelength $\lambda = v / f_0$, where v is the SAW velocity and f_0 is the transducer center frequency. For 128° YX-lithium niobate, the SAW velocity v in a semi-metallized region is 3940.5 m/s. For $f_0 = 1$ GHz, $\lambda = 3.941$ microns. Therefore, the finger width for single-electrode (i.e., $\lambda / 4$) fingers is 0.985 microns, and the finger width for double-electrode (i.e., $\lambda / 8$) fingers is 0.493 microns. Since the photolithography imposes a minimum linewidth requirement of 1 micron, the IDTs must be of the single-electrode type.

The input and output transducers of a SAW device consist of interdigital metal fingers connected to common bus bars. At each metal electrode edge, the acoustic surface wave encounters a discontinuity in the propagation medium. This discontinuity causes part of the surface wave to be reflected. The metal electrodes affect both the velocity of propagation of the surface acoustic wave as well as the characteristic impedance of the propagation medium. For single-electrode transducers, the reflections add in phase, whereas for double-electrode transducers, the reflections cancel at the center frequency of operation. The effect of these finger reflections is to introduce distortions in the frequency-domain and time-domain responses of the device. Examples of these distortions include ripples in the frequency-domain response and spurious signals in the time-domain response.

Both wideband and narrowband, single-electrode, phase-coded SAW designs were investigated in order to evaluate their relative performance. Wideband designs use less fingers in total than narrowband designs use. Therefore, the distortions caused by finger reflections are less pronounced for wideband designs. Wideband designs also use more fingers in the input IDT than narrowband designs use. Therefore, the input IDT match should be less sensitive to parasitic effects for wideband designs. However, wideband designs have alternating semi-metallized and non-metallized areas for the coded transducer, whereas narrowband designs do not. Therefore, the velocity of the surface acoustic wave is held more constant for narrowband designs than for

wideband designs. For single-electrode designs, the use of dummy electrodes in the non-metallized areas is not recommended, since these will exacerbate the finger reflection problem.

The mask layout information for the phase-coded SAW designs was generated using the PCS_{*n*} FORTRAN layout program (*n* indicates the application-specific version). This program uses the design parameters shown in Table 5.1.

Table 5.1. PCS_{*n*} layout program design parameters.

Parameter Name	Parameter Description
DX	$\lambda / 2$ finger spacing (microns)
W	acoustic aperture of IDT (wavelengths)
NFI	number of fingers in the input IDT
NFO ⁽¹⁾	number of fingers in the output IDT
NCYCLS ⁽¹⁾	number of carrier cycles per chip
NFT ⁽²⁾	number of fingers in each rung of the tapped output IDT
NTAP ⁽²⁾	number of taps in the output IDT
NFILT	number of matched filters in the filter bank
DHIDT	edge-to-edge horizontal distance between input and nearest output IDT (wavelengths)
DVIDT	edge-to-edge vertical distance between IDTs (wavelengths)
V ⁽³⁾	SAW velocity
TS ⁽³⁾	symbol time
CODE(I,J)	phase-coded sequences for each SAW filter

Fig. 5.4 shows the die layout for the phase-coded SAW matched filter bank. The die size is 2.5 mm X 7.5 mm. The matched filter input and output transducer patterns, bus bars and connecting pads were defined on the surface

¹ For a narrowband design.

² For a wideband design.

³ This parameter is used to compute the distance between output IDTs in a matched filter bank that uses differential coherent detection.

Figure 5.4. Phase-coded SAW matched filter bank die layout.

of the substrate using conventional photolithography techniques. The smallest dimension on the photomask is the interdigital finger width (approximately 1 micron at 1 GHz). The transducers were fabricated using four different metallization systems: (1) 400Å thick aluminum film, (2) 1000Å thick aluminum film, (3) 400Å thick gold film, and (4) 1000Å thick gold film. To facilitate metal film adhesion, a thin layer (approx. 100Å) of titanium was first deposited on the lithium niobate substrate.

The lo-Z buffers and multipliers shown in Fig. 4.5 have been integrated in a custom RF IC. The block diagram for the buffer-multiplier (BUFMULT) integrated circuit is shown in Fig. 5.5. It consists of six input buffer amplifiers (A1-A6), three multipliers (M1-M3) and three output buffer amplifiers (B1-B3). The purpose of the BUFMULT is to amplify and multiply the outputs from each SAW matched filter pair. The input amplifiers are transimpedance amplifiers with a transimpedance of 250 ohms and an input impedance of 10 ohms. The low input impedance is necessary to minimize multiple-transit reflections in the SAW device. The multipliers are four-quadrant bilinear tree multipliers with scaling constants of 20 V⁻¹. The output buffers have a gain of 0.5 V/V when driving off-chip to a 50 ohm load (25 ohms/side, differential).

Figure 5.5. Buffer-multiplier (BUFMULT) RF integrated circuit.

Figure 5.6. Differential coherent detector designer test (DT) module.

Figure 5.7. Differential coherent detector multi-chip module (MCM).

Fig. 5.6 presents a scanned image of the differential coherent detector DT module. The DT module implements a bandpass version of the GL receiver front-end shown in Fig. 4.4. The DT module is comprised of the SAW matched filter and BUFMULT dies mounted in a 14-pin gold-plated DIP. A microwave chip capacitor is used to filter the power supply lines to the BUFMULT die. The size of the SAW die is 2.5 mm X 7.5 mm and the size of the BUFMULT die is 1.5 mm X 2.5 mm. Bond wires are used to interconnect the two dies, the chip capacitor and the pins of the gold-plated DIP. One input to the multiplier is taken from the undelayed matched filter output corresponding to code 01111. The other input to the multiplier is taken from the delayed matched filter output corresponding to code 10001. The SAW die was fabricated using Nortel's SAW fabrication process. The BUFMULT die was fabricated using Nortel's NT25 high-speed silicon bipolar process.

Fig. 5.7 depicts the differential coherent detector multi-chip module. This MCM implements the circuitry shown in Fig 5.5, except for the low pass

filters, which are external to the module. The MCM consists of a thin-film alumina substrate carrier mounted in a 48-pin gold-plated dip. The SAW and BUFMULT dies, microwave chip capacitors and other surface-mount components are attached to the alumina substrate using epoxy. In a differential coherent detection scheme, the output of one matched filter is multiplied by the delayed output of another matched filter. Since the signals at each matched filter output are bandpass signals (i.e., modulated on a 1 GHz carrier), it is necessary to ensure an absolute time delay accuracy of about ± 100 ps ($\pm 10\%$ of the carrier period). Measurements of the time delay between the undelayed and delayed correlated output signals from a matched filter pair, for several devices on two different wafers, indicated a nominal value of 24.8 ns and a delay spread of about ± 50 ps. It was, therefore, decided to trim the symbol period delay provided by the matched filter SAW by including surface-mount delay lines (200 ps nominal) on the substrate. Since the delay lines are thin-film 50-ohm transmission lines, a matching network is required to match the input impedance ($10\Omega + j10\Omega$ nominal) of the lo-Z buffers (on the BUFMULT die) to the 50 ohm impedance of the delay lines. The matching network employs quarter-wave transmission line transformers and 50 ohm transmission line stubs. The characteristic impedance of the quarter-wave transformers is approximately 28 ohms. The quarter-wave transformers and 50 ohm stubs are defined as patterned coplanar waveguide on the 10-mil alumina substrate. For the quarter-wave transformers, the center conductor width is 600 microns and the gap is 200 microns. For the 50 ohm transmission line stubs, the center conductor width is 100 microns and the gap is 50 microns. Via holes are employed for good ground conductivity between the top and bottom gold layers of the substrate. Bond wires are used to make interconnections between the dies, chip components, transmission lines and package pins.

Chapter 6

Measurements and Comparison to Theory

The commutation signaling differential coherent detector uses phase-coded SAW matched filters to perform the signal processing function of correlation. The SAW devices were fabricated using three different metal systems. This chapter presents measurement results for both wideband and narrowband phase-coded SAW correlators, and discusses the effect of metalization on device performance. SAW device time-domain and frequency-domain measurements are compared against theoretical predictions obtained using the Tancrrell and Holland delta-function model [14]. A discussion of second-order effects is included. This chapter also presents measurement results for the commutation signaling differential coherent detector.

The impulse response of a matched filter is a time-reversed replica of the expected received signal (e.g., commutation waveform). Matched filters perform the mathematical operation of correlation. With bandwidth expanding signals, matched filters are pulse compression filters since the output signal consists of a time-compressed main lobe surrounded by leading and trailing sidelobes. Minimizing these sidelobes is a common design criterion for matched filters used with bandwidth expanding signals. Hence, their performance is often evaluated on the basis of time-domain measurements. Time-domain measurements for wideband and narrowband SAW devices are presented in sections 6.1, 6.2 and 6.3. Section 6.1 discusses impulse response measurements, section 6.2 discusses autocorrelation response measurements, and section 6.3 discusses cross-correlation response measurements. Section 6.4 analyzes the performance of these devices on the basis of network analyzer time-domain and frequency-domain measurements. Section 6.5 presents measurement results for the commutation signaling differential coherent detector.

6.1 Impulse Response Measurements

The simplest method for evaluating the time-domain performance of SAW filters is to observe their impulse response. Fig. 6.1 shows impulse response measurements for both a wideband and a narrowband phase-coded SAW filter. These filters generate three successive 5-chip codewords. The chip duration is 5 ns. These measurements were made in a 50 ohm system using a wide-bandwidth digital sampling oscilloscope. The impulse response for each filter was generated by exciting the input IDT with a 500 ps wide (one half cycle at 1 GHz) 5V impulse (applied via a 3-dB attenuator). The impulse response measurements exhibit an amplitude "droop" over the 75 ns impulse response duration. The "droop" is about 5 dB for the wideband device and about 10 dB for the narrowband device. The narrowband device also has larger insertion loss due to greater electrical mismatch.

Figure 6.1. Comparison of wideband and narrowband phase-coded SAW impulse response measurements. Coded sequence: 100010111101101; 1000Å aluminum metallization.

It is instructive to compare Fig. 6.1 with Fig. 4 in the paper by Carr *et al.* [15]. The latter figure shows the impulse response of a phase-coded SAW tapped correlator operating at a chip rate of 200 Mc/s at a center frequency of 1 GHz. This impulse response shows a "droop" of 3.5 dB over the 75 ns impulse response duration. The authors attribute this loss to the scattering of the surface acoustic wave into volume waves at the discontinuities created by the thin-film metal electrodes.

Fig. 6.2 shows impulse response measurements for wideband and narrowband phase-coded SAW filters that generate one 5-chip codeword. The sample number interval corresponds to a time span of about 50 ns (10 chip periods). It is seen that the amplitude "droop" varies as a function of the particular code sequence used. The worst case occurs for those codes that have the largest number of in-phase fingers. Narrowband devices exhibit more amplitude "droop" than wideband devices.

Figure 6.2. Comparison of wideband and narrowband phase-coded SAW impulse response measurements. 1000Å aluminum metallization.

The operation of a 2.4 GHz SAW correlator fabricated on an AlN/Al₂O₃ substrate is described in the paper by Nakase *et al.* [16]. This correlator operates at a chip rate of 22 Mc/s. Fig. 2 in that paper shows the impulse response of the correlator (measured using a network analyzer). The amplitude "droop" is 15 dB over the 500 ns impulse response duration.

6.2 Autocorrelation Response Measurements

A block diagram of the measurement system used to evaluate matched filter time-domain performance is shown in Fig. 6.3. The measurement set-up consists of the following: (1) a high-speed pulse generator, (2) the SAW code generator module, (3) the SAW matched filter module (DUT), and (4) a wide-bandwidth digital sampling oscilloscope. The measurement equipment provides 50 ohm source and load terminations. The SAW code generator module produces the phase-coded sequence needed for driving the SAW matched filter module. Both SAW modules employ microwave amplifiers to compensate for the insertion loss of the individual SAW devices. Due to the wide-bandwidth nature of the signals employed, no attempt was made to match the SAW devices. The pulse generator produces a 500 ps wide pulse to stimulate the impulse response (representing one 5-chip codeword) of the code generator SAW filter. Time-domain measurements are captured using a wide-bandwidth digital sampling oscilloscope.

Fig.'s 6.4-6.6 show theoretical and experimental autocorrelation response plots for *wideband* SAW filters matched to commutation signaling codes 01101, 01111 and 10001 respectively. These codes were selected using the code search technique described in [10]. The dotted curve in the theoretical figure window represents the envelope of the ideal autocorrelation response. The theoretical plots were created using MATLAB® computer simulations based on the Tancrell and Holland delta-function model. Comparison of the theoretical and experimental waveforms in Fig.'s 6.4-6.6 shows reasonable agreement between measured and predicted responses. The differences between

Figure 6.3. Measurement system for evaluating SAW matched filter time-domain performance.

the observed and theoretical waveforms are due to second-order effects that are not taken into account by the delta-function model.

Table 6.1 provides a comparison of measured and theoretical autocorrelation response peak-to-maximum-sidelobe ratios for the wideband devices. The peak-to-maximum-sidelobe ratio is defined as the ratio of the main lobe amplitude to the maximum sidelobe amplitude in dB. The deviation from theoretical varies by 1.4-3.8 dB. This is attributed mainly to acoustic mismatch (i.e., finger reflections) at the metal electrode edges [17]. The input and output transducers of a SAW device consist of interdigital metal fingers connected to common bus bars. At each metal electrode edge, the acoustic surface wave encounters a discontinuity in the propagation medium. This discontinuity causes part of the surface wave to be reflected. The metal electrodes affect both the velocity of propagation of the surface acoustic wave as well as the characteristic impedance of the propagation medium. The effect of these finger reflections is to introduce distortions in the time-domain response of the device. As a consequence, the SAW filter is no longer perfectly matched to the received waveform. Therefore, the experimentally measured matched filter output deviates from the theoretical autocorrelation function.

Table 6.1. Comparison of peak-to-maximum-sidelobe ratios (wideband devices).

Matched filter code	Measured peak-to-max.-sidelobe ratio	Theoretical peak-to-max.-sidelobe ratio	Deviation	Reference
10110	6.1 dB	7.5 dB	1.4 dB	Fig. 6.4
11110	10.6 dB	14.4 dB	3.8 dB	Fig. 6.5
10001	4.9 dB	7.7 dB	2.8 dB	Fig. 6.6

Figure 6.4. Comparison of theoretical and measured SAW matched filter autocorrelation responses. Matched filter code: 10110, wideband tapped IDT; 1000Å aluminum metallization.

Figure 6.5. Comparison of theoretical and measured SAW matched filter autocorrelation responses. Matched filter code: 11110, wideband tapped IDT; 1000Å aluminum metallization.

Figure 6.6. Comparison of theoretical and measured SAW matched filter autocorrelation responses. Matched filter code: 10001, wideband tapped IDT; 1000Å aluminum metallization.

Fig.'s 6.7-6.9 show theoretical and experimental autocorrelation response plots for *narrowband* SAW filters matched to commutation signaling codes 01101, 01111 and 10001 respectively. The dotted curve in the theoretical figure window represents the envelope of the ideal autocorrelation response. The difference between the observed and predicted waveforms is more pronounced than in the case of the wideband devices. This is explained as follows. The narrowband devices use more fingers in the coded transducer than do the wideband devices, and have more fingers in total. Since the time-domain response of these filters is affected mainly by finger reflections, the effects of acoustic mismatch are more pronounced in the case of the narrowband devices.

Figure 6.7. Comparison of theoretical and measured SAW matched filter autocorrelation responses. Matched filter code: 10110, narrowband coded IDT; 1000Å aluminum metallization.

Figure 6.8. Comparison of theoretical and measured SAW matched filter autocorrelation responses. Matched filter code: 11110, narrowband coded IDT; 1000Å aluminum metallization.

Figure 6.9. Comparison of theoretical and measured SAW matched filter autocorrelation responses. Matched filter code: 10001, narrowband coded IDT; 1000Å aluminum metallization.

Table 6.2 provides a comparison of measured and theoretical autocorrelation response peak-to-maximum-sidelobe ratios for the narrowband devices. The deviation from theoretical varies by 2.5-6.5 dB. This is 1.1-2.7 dB worse than for the wideband devices.

Table 6.2. Comparison of peak-to-maximum-sidelobe ratios (narrowband devices).

Matched filter code	Measured peak-to-max.-sidelobe ratio	Theoretical peak-to-max.-sidelobe ratio	Deviation	Reference
10110	5.0 dB	7.5 dB	2.5 dB	Fig. 6.7
11110	7.5 dB	14.0 dB	6.5 dB	Fig. 6.8
10001	2.1 dB	7.7 dB	5.6 dB	Fig. 6.9

In order to understand the differences between the theoretical and measured output responses in Fig.'s 6.4-6.9, it is necessary to analyze actual SAW device impulse response measurements. These measurements are shown in Fig.'s 6.10-6.15 together with measured and calculated autocorrelation responses. The calculated autocorrelation response was obtained by mathematically convolving the measured impulse response data for the code generator and matched filter SAWs. It is seen that the calculated response is in good agreement with the measured response for all devices. Thus, by understanding the cause of the distortions in the measured impulse responses, it is possible to understand the differences between the theoretical and measured output responses in Fig.'s 6.4-6.9.

The impulse response measurements exhibit two main types of distortion: (1) a definite taper (amplitude droop) over the duration of the impulse response; and (2) significant trailing energy after the main pulse. The amplitude droop is more pronounced for the narrowband devices than for the wideband devices. In [17], similar types of distortion were observed for the impulse responses of phase-coded SAW correlators fabricated using aluminum metallization on lithium niobate. The authors concluded that the major contribution to the observed distortions was acoustic mismatch at the electrode discontinuities (i.e., finger reflections). Most other effects, including electro-acoustic regeneration, were shown to be small compared to those due to finger reflections. The narrowband devices use more fingers in the coded transducer than the wideband devices use, and have more fingers in total. Therefore, the effects of acoustic mismatch are more pronounced in the case of the narrowband devices.

Figure 6.10. Comparison of calculated (from impulse response data) and measured SAW matched filter autocorrelation responses. Matched filter code: 10110, wideband tapped IDT; 1000Å aluminum metallization.

Figure 6.11. Comparison of calculated (from impulse response data) and measured SAW matched filter autocorrelation responses. Matched filter code: 11110, wideband tapped IDT; 1000Å aluminum metallization.

Figure 6.12. Comparison of calculated (from impulse response data) and measured SAW matched filter autocorrelation responses. Matched filter code: 10001, wideband tapped IDT; 1000Å aluminum metallization.

Figure 6.13. Comparison of calculated (from impulse response data) and measured SAW matched filter autocorrelation responses. Matched filter code: 10110, narrowband coded IDT; 1000Å aluminum metallization.

Figure 6.14. Comparison of calculated (from impulse response data) and measured SAW matched filter autocorrelation responses. Matched filter code: 11110, narrowband coded IDT; 1000Å aluminum metallization.

Figure 6.15. Comparison of calculated (from impulse response data) and measured SAW matched filter autocorrelation responses. Matched filter code: 10001, narrowband coded IDT; 1000Å aluminum metallization.

The problem of acoustic reflections at interdigital electrode edges can be solved by using an IDT structure consisting of double-electrodes. However, this requires higher photolithography resolution and, hence, imposes an upper limit on the frequency range of the SAW devices. Another approach to suppressing acoustic reflections in single-electrode IDT structures is based on the observation that, for certain types of metal/substrate combinations, an anomalous reflection phenomenon exists. It has been shown theoretically [18] that, for some metal/substrate combinations, the magnitude of the reflection coefficient, as a function of increasing metal-thickness-to-wavelength ratio (h/λ), first decreases and then increases after reaching a minimum. By choosing the appropriate h/λ , internal IDT reflections are minimized.

For the *short-circuited* metal-strip grating, the reflection coefficient of a single strip in the grating can be expressed as [18]

$$\begin{aligned} r_s &= \left[\frac{C'_{11}}{C_{11}} I_1(\nu', \nu) + \frac{\rho'}{\rho} I_2(\nu) \right] f_1(\eta\pi)(h/\lambda) + f_3(\eta\pi) \frac{K^2}{2} \\ &= R_m \frac{h}{\lambda} + R_e \end{aligned} \quad (6.1)$$

where the primed quantities refer to the strip medium (metal) and the unprimed quantities refer to the substrate medium. h is the strip thickness and λ is the wavelength at the Bragg (center) frequency. Equation (6.1) is an *approximate* expression since it was derived using the concept of an "equivalent isotropic medium". In [18], an "equivalent isotropic medium" is defined in the following sense: (1) it possesses the same mass density as the actual substrate; (2) its Poisson ratio is chosen by fitting the reflectivity of a groove grating with the experimental results; and (3) it possesses the same piezoelectric tensor and dielectric tensor as the actual substrate. In equation (6.1), ν denotes the equivalent Poisson ratio, C_{11} denotes the equivalent elastic constant and ρ denotes the mass density of the medium. K^2 denotes the electromechanical coupling coefficient of the piezoelectric substrate. The functions $f_1(\eta\pi)$ and $f_3(\eta\pi)$ are geometric factors that depend on the strip metallization ratio η and on the Bragg frequency at which the grating is operating. For a 50% metallization ratio (i.e., $\eta = 0.5$), $f_1(\eta\pi) = 1$ [18]. At the first Bragg frequency (i.e., $f_0 = v/\lambda$), $f_3(\eta\pi)$ is given by [18]

$$f_3(\eta\pi) = -\frac{\pi}{2} \left\{ \cos(\eta\pi) + \frac{P_{0.5}[-\cos(\eta\pi)]}{P_{-0.5}[-\cos(\eta\pi)]} \right\} \quad (6.2)$$

where $P_s(x)$ are Legendre functions of fractional order s . For $\eta = 0.5$, $f_3(\eta\pi) = -0.71$.

An isotropic medium is characterized by three elastic constants C_{11} , C_{12} and C_{44} , of which two are independent (i.e., $C_{12} = C_{11} - 2C_{44}$) [19]. These are often taken to be the Lamé constants, defined by

$$\begin{aligned} \lambda &= C_{12} \\ \mu &= C_{44} \end{aligned} \quad (6.3)$$

In terms of the Lamé constants, Poisson's ratio ν for an isotropic medium is defined as

$$\nu = \frac{\lambda}{2(\lambda + \mu)} \quad (6.4)$$

For an isotropic medium, the velocity of a compressional (i.e., longitudinal) bulk wave propagating in the z -direction is given by [19]

$$v_l = \sqrt{\frac{C_{11}}{\rho}} \quad (6.5)$$

The velocities of the corresponding x -polarized and y -polarized, z -propagating shear waves are given by [19]

$$v_s = \sqrt{\frac{C_{44}}{\rho}} \quad (6.6)$$

Since the medium is isotropic, these solutions are valid for all propagation directions. For an anisotropic medium, the longitudinal and shear wave velocities will, in general, depend on the propagation direction. These velocities can be determined by solving the Christoffel equation for the particular anisotropic medium [19]. In general, numerical computation is required in order to obtain a complete solution.

In [20], the longitudinal and shear bulk wave velocities have been tabulated for a variety of different crystalline materials, including lithium niobate. For X, 41.5-plane LiNbO_3 (i.e., 131° YX-lithium niobate [21]), the longitudinal

and slow shear wave velocities are 6,575 m/s and 4,080 m/s respectively. Since this cut is close to the 128°-rotated cut [22], these velocities are used to estimate the bulk wave velocities for 128° YX-lithium niobate. Substituting the longitudinal wave velocity into equation (6.5) yields the equivalent elastic constant C_{11} as $20.3 \cdot 10^{10}$ N/m². Substituting the slow shear wave velocity into equation (6.6) yields the equivalent elastic constant C_{44} as $7.82 \cdot 10^{10}$ N/m². The equivalent Poisson ratio is then obtained from equations (6.3) and (6.4) using the relation $C_{12} = C_{11} - 2C_{44}$. The equivalent Poisson ratio and elastic constants for the substrate material (128° YX-lithium niobate) are shown in Table 6.3. The Poisson ratio and elastic constants for two types of strip materials (polycrystal gold and polycrystal aluminum) are also shown in Table 6.3. The material densities and strip material constants were obtained from [19].

Table 6.3. Equivalent Poisson ratios and elastic constants for substrate and strip materials.

Material	ν	C_{11} (10^{10} N/m ²)	C_{44} (10^{10} N/m ²)	ρ (kg/m ³)
128° YX-LiNbO ₃	0.187	20.3	7.82	4,700
gold	0.420	20.7	2.85	19,300
aluminum	0.355	11.1	2.50	2,700

In equation (6.1), $R_m(h/\lambda)$ is the contribution due to mechanical (mass and elastic stress) loading, and R_e is the contribution due to piezoelectric shorting. R_e is always negative. R_m , however, can be either positive or negative, depending on the material characteristics of the metal and substrate. Hence, a minimum, even zero reflection coefficient could be achieved by optimizing h/λ . For 128° YX-lithium niobate, $K^2 = 0.053$. From [18], for gold metallization on 128° YX-lithium niobate, $I_1(\nu', \nu) = -1.40$ and $I_2(\nu) = 0.53$. For aluminum metallization on 128° YX-lithium niobate, $I_1(\nu', \nu) = -1.47$ and $I_2(\nu) = 0.53$. Substituting these values into equation (6.1) gives the following theoretical values for R_m and R_e .

Table 6.4. Coefficients R_m and R_e in equation (6.1).

Strip Material	R_m	R_e
gold	0.75	-0.019
aluminum	-0.50	-0.019

Equation (6.1) predicts that, for gold metallization on 128° YX-lithium niobate, a strip thickness of $1000\text{\AA}$ should theoretically give a reflection coefficient close to zero for a short-circuited metal strip. For aluminum metallization on 128° YX-lithium niobate, equation (6.1) predicts that it is theoretically not possible to obtain a zero reflection coefficient with a 50% strip metallization ratio. The minimum reflection coefficient is obtained by minimizing the metal film thickness. Equation (6.1) is an approximate expression for the following reasons: (1) it was derived using the concept of an "equivalent isotropic medium", and (2) it neglects the effect of piezoelectric regeneration since it assumes a short-circuited grating. In [23], the theory presented by Chen and Haus [18] is extended to the practical case of an anisotropic medium. The resulting expression for R_m is no longer an analytic function of the material constants and needs to be solved by numerical methods.

In [23] it is shown that R_m is a complex number. This means that the contribution to the reflection coefficient due to mechanical loading is not exactly in-phase with the contribution due to piezoelectric shorting. Consequently, the minimum value of the magnitude of the reflection coefficient is not quite zero. Using the values for R_m and R_e shown in Table I in the article by Chen and Haus [18], equation (6.1) predicts that the reflection coefficient is zero when $h \setminus \lambda = 1.40\%$, for gold metallization on YZ-lithium niobate. In [23], the value of $h \setminus \lambda$ that gives a minimum reflection coefficient is shown to be 0.73% , for gold metallization on YZ-lithium niobate. The latter value was determined by fitting experimental data to the expression for R_m for an anisotropic medium. It also takes into account piezoelectric regeneration caused by transducer loading.

In order to experimentally investigate the effect of different metallization types on matched filter performance, devices were fabricated using $400\text{\AA}$ and $1000\text{\AA}$ thick aluminum film and $400\text{\AA}$ and $1000\text{\AA}$ thick gold film. It should be noted that SAW process variations and cleanliness both become more critical as the film thickness decreases. The plot on the next page shows the autocorrelation response of a wideband SAW filter matched to commutation signaling code 10001. This device was fabricated using $1000\text{\AA}$ thick gold film. The measured response deviates significantly from the theoretical autocorrelation response shown in Fig. 6.6.

Figure 6.16. Effect of different metallization types on SAW matched filter autocorrelation response. Matched filter code: 10001; wideband tapped output IDT. Upper figure: 400Å aluminum metallization. Middle figure: 1000Å aluminum metallization. Lower figure: 400Å gold metallization.

Fig. 6.16 presents the effect of the other three metallization types on the autocorrelation response of a wideband SAW filter matched to commutation signaling code 10001. The responses shown in the upper, middle and lower figures correspond to the following metallization: (upper) 400Å aluminum, (middle) 1000Å aluminum, and (lower) 400Å gold. The measured peak-to-left-sidelobe and peak-to-right-sidelobe ratios are also indicated. For the three devices, there was little difference in the value of the peak-to-left-sidelobe ratio. The distinguishing difference among the measured responses is the appearance of the trailing sidelobes. Distortion of the trailing sidelobes is typical of phase-coded SAW correlators [17], [24]. As discussed previously, acoustic reflections at the metal electrode edges due to mechanical loading and piezo-electric shorting are the major cause of this distortion. Since all three devices

use the same interdigital transducer pattern, the effect of piezoelectric shorting on the appearance of the trailing sidelobes should be similar. It is concluded, therefore, that the distinguishing differences among the measured responses of the three devices result from mechanical loading. The 400Å aluminum device appears to have the best overall performance. The performance of the 400Å gold device is not too dissimilar from the aluminum devices. It is believed that its performance could be improved by optimizing h / λ [23].

6.3 Cross-Correlation Response Measurements

Fig.'s 6.17-6.22 present theoretical and experimental cross-correlation response plots for wideband SAW correlators matched to commutation signaling codes 01101, 01111 and 10001 respectively. The dotted curves in these figures represent the autocorrelation responses. Experimental results were obtained using the test set-up shown in Fig 6.3. As discussed above, the differences between the observed and theoretical waveforms are mainly due to acoustic reflections at the interdigital electrode edges. As expected, the cross-correlation responses in Fig.'s 6.17, 6.18 & 6.19 are time-reversed replicas of the cross-correlation responses in Fig.'s 6.20, 6.21 & 6.22 respectively.

Table 6.5 provides a comparison of measured and theoretical peak-autocorrelation-to-crosscorrelation-ratios for the wideband devices. This ratio is defined as the ratio, in dB, of the autocorrelation response to the cross-correlation response at the instant when the autocorrelation response is maximum. The deviation from theoretical varies by 0.1-2.0 dB as compared to 1.4-3.8 dB for the autocorrelation measurements (Table 6.1). This can be explained as follows. As the acoustic wave propagates along the device, some of its energy is lost due to acoustic reflections. Thus, the mainlobe and right sidelobes are relatively more attenuated than the left sidelobes. The autocorrelation peak-to-maximum-sidelobe ratio is calculated with respect to the left sidelobes, whereas the peak-autocorrelation-to-crosscorrelation-ratio is calculated at the same time instant. This accounts for the differences.

Table 6.5. Comparison of peak-autocorrelation-to-crosscorrelation-ratios (wideband devices).

Matched filter code	Transmitted code	Measured peak-auto-to-cross-correlation ratio	Theoretical peak-auto-to-cross-correlation ratio	Deviation	Ref.
10110	01111	3.5 dB	4.0 dB	0.5 dB	Fig. 6.17
10110	10001	11.2 dB	12.4 dB	1.2 dB	Fig. 6.18
11110	10001	5.2 dB	4.5 dB	0.7 dB	Fig. 6.19
11110	01101	2.9 dB	4.9 dB	2.0 dB	Fig. 6.20
10001	01101	11.8 dB	12.9 dB	1.1 dB	Fig. 6.21
10001	01111	4.0 dB	4.1 dB	0.1 dB	Fig. 6.22

Figure 6.17. Comparison of theoretical and measured SAW matched filter cross-correlation responses. Transmitted code: 01111; matched filter code: 10110, wideband tapped IDTs; 1000Å aluminum metallization.

Figure 6.18. Comparison of theoretical and measured SAW matched filter cross-correlation responses. Transmitted code: 10001; matched filter code: 10110, wideband tapped IDTs; 1000Å aluminum metallization.

Figure 6.19. Comparison of theoretical and measured SAW matched filter cross-correlation responses. Transmitted code: 10001; matched filter code: 11110, wideband tapped IDTs; 1000Å aluminum metallization.

Figure 6.20. Comparison of theoretical and measured SAW matched filter cross-correlation responses. Transmitted code: 01101; matched filter code: 11110, wideband tapped IDTs; 1000Å aluminum metallization.

Figure 6.21. Comparison of theoretical and measured SAW matched filter cross-correlation responses. Transmitted code: 01101; matched filter code: 10001, wideband tapped IDTs; 1000Å aluminum metallization.

Figure 6.22. Comparison of theoretical and measured SAW matched filter cross-correlation responses. Transmitted code: 01111; matched filter code: 10001, wideband tapped IDTs; 1000Å aluminum metallization.

6.4 Network Analyzer Measurements

Fig. 6.24 presents time-domain measurements for wideband phase-coded SAW filters. These measurements were made in a 50 ohm system using the test set-up shown in Fig. 6.23. Time-domain measurements were obtained using the transform option on the vector network analyzer (VNA). These measurements show three significant second-order effects: (1) electromagnetic feed-through (EMFT), (2) finger reflections, and (3) triple-transit interference (TTI). Marker 1 is the reference, and is positioned at the peak of the EMFT pulse. Marker 2 is positioned at the center of the main pulse response. Marker 3 is positioned at the center of the triple-transit response (i.e., the time delay to marker 3 is three times the time delay to marker 2).

EMFT is a parasitic effect caused by electromagnetic signal leakage from the input to the output of the device. This signal leakage originates from four possible sources: (1) resistive coupling, (2) capacitive coupling, (3) magnetic coupling and (4) radiative coupling. It is very dependent on package design and ground layout. For the measurements shown in Fig. 6.24, the EMFT is only about 17 dB below the main signal. This is explained as follows. The SAW devices were mounted in 14-pin gold-plated DIPs. The ground leads for the input and output transducers were bonded to the common package ground (i.e., the input and output IDT ground returns were not isolated). In

Figure 6.23. Network analyzer measurement test set-up.

(a)

(b)

(c)

Figure 6.24. Phase-coded SAW time-domain measurements. Wideband tapped output IDT; 1000Å aluminum metallization. (a) Coded sequence: 01101. (b) Coded sequence: 01111. (c) Coded sequence: 10001.

general, EMFT becomes worse as the operating frequency increases. Therefore, due to the high operating frequency (1 GHz) and lack of ground isolation, the EMFT is high. As discussed previously, finger reflections are caused by acoustic mismatch at the interdigital electrode edges. Each electrode reflects a part of the incident acoustic wave. For single-electrode fingers, the reflected waves add coherently since the electrode spacing equals half the wavelength. Hence, the overall effect is significant, as shown in Fig. 6.24. Triple-transit is an unwanted output signal due to surface waves traversing the device three times. TTI is minimized by deliberately mismatching the source and load. Fig. 6.24 shows that the triple-transit signal is about 25-30 dB below the main signal. This can be reduced by terminating the output transducer in a low-impedance load (e.g., connecting the SAW output to the lo-Z input on the BUFMULT RF IC).

Fig. 6.25 presents the effect of the four metallization types on the frequency-domain response of a SAW filter that generates commutation signaling code 10001. Measured (solid curve), theoretical (dashed curve) and calculated (dashdot curve) frequency response plots are shown for the 400Å aluminum device (a), the 1000Å aluminum device (b), the 400Å gold device (c), and the 1000Å gold device (d). The theoretical and calculated frequency responses were obtained by using an FFT algorithm to transform the delta-function model and measured impulse response coefficients respectively. The experimental frequency response data (601 points covering a 600 MHz frequency span) was obtained by means of a VNA. The theoretical frequency response curves have been shifted to compensate for the center frequency deviation of the actual SAW devices. As expected, due to the mass loading effect, the gold devices operate at a lower center frequency than the aluminum devices, and the 1000Å devices operate at a lower center frequency than the 400Å devices. The frequency offset ranges from 4 MHz for the 400Å aluminum device to 132 MHz for the 1000Å gold device. The deviation between the measured and theoretical frequency response curves is due to a combination of second-order effects (electromagnetic feed-through, finger reflections and triple-transit interference). The rms magnitude error (i.e., the difference between the measured and the shifted theoretical frequency response) is 0.083 for the 400Å aluminum device, 0.093 for the 1000Å

aluminum device, 0.147 for the 400Å gold device and 0.213 for the 1000Å gold device.

Figure 6.25. Comparison of measured (solid curve), theoretical (dashed curve) and calculated (dashdot curve) SAW filter frequency responses. SAW filter code: 10001; wideband tapped output IDT. (a) 400Å aluminum metallization. (b) 1000Å aluminum metallization. (c) 400Å gold metallization. (d) 1000Å gold metallization.

6.5 Differential Coherent Detector Measurements

An important benefit of differentially encoded binary commutation signaling is that it reduces the sidelobe level of the signal applied to the RAKE receiver. Fig. 6.27 presents an experimental confirmation of the sidelobe reduction that occurs at the differential coherent detector output as a result of the multiplication. Three waveforms are shown in Fig. 6.27. The top waveform corresponds to the delayed output of matched filter #1 (matched to commutation codeword 10001). The middle waveform corresponds to the output of matched filter #2 (matched to commutation codeword 01111). The bottom waveform corresponds to the low-pass filtered output of the differential coherent detector. The measurements in Fig. 6.27 were obtained using the test set-up shown in Fig. 6.26. A signal consisting of a continuous repetition of the three commutation codewords 100010111101101 was applied to the input of the differential coherent detector DT module. The resulting peaks at the differential coherent detector output are spaced 75 ns apart (3 symbol periods).

Fig. 6.27 shows that the signal at the output of the differential coherent detector contains smaller sidelobes when compared against the signals present at the output of each matched filter. This is explained as follows. The differential coherent detector output is obtained by multiplying the delayed output of matched filter #1 by the output of matched filter #2. The resultant signal is then low-pass filtered. The signals at each matched filter output have their mainlobes located at the same position in time. The sidelobes, however, are spread over different locations in time. Therefore, after multiplication, the mainlobes are enhanced and the sidelobes are reduced. Differences in the matched filters due to second-order effects and fabrication process variations reduce the detection peaks and enhance the sidelobes.

Figure 6.26. Differential coherent detector DT module measurement test set-up.

Figure 6.27. Experimental measurements showing sidelobe reduction at output of differential coherent detector designer test (DT) module.

Fig. 6.28 shows the waveforms present at each low-pass filtered output of the differential coherent detector multi-chip module (Fig. 5.7) when the input is driven by a signal consisting of a continuous repetition of the three commutation codewords 100010111101101. Each output peak corresponds to one demodulated data symbol. Each output waveform is obtained by multiplying the delayed output of one matched filter by the undelayed output of another matched filter, as summarized in Table 6.6. The peaks in each waveform are spaced 75 ns apart (3 symbol periods). The peaks between adjacent waveforms are spaced 25 ns apart (one symbol period). The sidelobes between peaks are reduced by the multiplication effect, but are still significant because of the very short spreading sequences employed (5 chips/bit). The middle waveform in Fig. 6.28 corresponds to the lower waveform in Fig. 6.27. These

Figure 6.28. Differential coherent detector multi-chip module (MCM) outputs.

two waveforms are similar in appearance. The differences between these two measured responses are explained as follows. For the DT module, only two SAW matched filter outputs were bonded to the BUFMULT chip, whereas for the MCM, all six SAW outputs were bonded. Therefore, the crosstalk between SAW outputs is less for the DT module than for the MCM. For the DT module, the SAW outputs were connected directly the BUFMULT chip ($10\ \Omega$ termination), whereas for the MCM, the SAW outputs were connected to the BUFMULT chip by means of a $50\ \Omega$ matching network. As a consequence, finger reflections are reduced in the case of the DT module since the SAW outputs are terminated in a lower impedance. The matching network also introduces some group delay distortion. These differences account for the differences in the observed responses.

Table 6.6. Differential coherent detector MCM input/output relationships.

Output wave- form reference (Fig. 6.28)	Matched filter code (delayed input)	Matched filter code (undelayed input)	Pk.-to-pk. amplitude (mV)
top	01101	10001	240
middle	10001	01111	181
bottom	01111	01101	114

In Fig 6.28, the signal amplitudes have been normalized in order to make it easier to compare waveforms. Actual peak-to-peak amplitudes are shown in Table 6.6. There is a 2.5 dB peak-to-peak amplitude variation between the top and middle waveforms in Fig. 6.28, and a 6.5 dB variation between the top and bottom waveforms. The reasons for these variations are explained below.

The plot on the next page shows the output of the SAW code generator module when excited by a single impulse. The code generator module generates three successive 5-chip codewords (100010111101101). Three types of impairments are evident in the plot: (1) a significant EMFT component; (2) an amplitude “droop” of about 5 dB over the 75 ns impulse response duration; and (3) a spurious response following the main pulse due to finger and multiple-transit reflections. The EMFT pulse occurs about 21 ns before the start of the main code sequence. It will, therefore, affect code 01101 of the previous code sequence the most. The amplitude “droop” will affect codeword 10001 the least and codeword 01101 the most. The spurious time-domain response will affect all three codewords of the next code sequence, although it appears that it will affect codeword 01111 more than codeword 10001.

The most attenuated signal in Fig 6.28 is the bottom waveform. This output is obtained by multiplying delayed codeword 01111 by undelayed codeword 01101. When considering the three codeword pair combinations shown in Table 6.6, this pair is the most affected by amplitude “droop”. Also, codeword 01101 is the most affected by EMFT, while codeword 01111 appears to be the most affected by spurious time-domain. It is concluded that these effects, together with SAW filter insertion loss variations, crosstalk between

PHASE-CODED SAW IMPULSE RESPONSE
 SHOWING EMFT
 CODED SEQUENCE: 10001011101101
 1000 Å ALUMINUM METALLIZATION

matched filter outputs, active device gain variations, and matching network group delay variations account for the amplitude differences shown in Table 6.6. The greatest improvement can be obtained by reducing EMFT and amplitude “droop” and time domain spuri due to finger reflections. Furthermore, it would be beneficial to design the group delay of the SAW filter so that the EMFT pulse occurs at the boundary between two codeword transitions. A different choice of codewords may also help. For example, codewords with long runs of 0’s or 1’s probably should be avoided.

As discussed previously, surface-mount delay lines are included on the MCM substrate in order to trim the symbol period delay provided by the matched filter SAW. Since the delay lines are thin-film 50-ohm transmission lines, a matching network is required to match the input impedance ($10\Omega + j10\Omega$ nominal) of the lo-Z buffers (on the BUFMULT die) to the 50 ohm impedance of the delay lines. The matching network employs quarter-wave transmission line transformers and 50 ohm transmission line stubs. These have been integrated on the thin-film alumina substrate. Wafer probing of partially assembled substrates was employed in order to verify the matching network design. Fig. 6.29 presents the results of the wafer probe/network analyzer measurements. S-parameter data (S_{11}) is presented in Smith chart and in graph (return loss) format at two different probe locations: (a) at the input to a quarter-wave matching transformer, and (b) at the input to a surface-mount delay line. It is seen that a good input match exists at the delay line input. The input match at the quarter-wave transformer input is not as good, but is still reasonable over a fairly large bandwidth.

(a)

(b)

Figure 6.29. Wafer probe/network analyzer S_{11} measurements. (a) at quarter-wave matching transformer input. (b) at surface-mount delay line input.

Chapter 7

Summary - Part I

Part I of this thesis has described a SAW-based commutation signaling modem for broadband indoor wireless communication. This system employs binary differential encoding, bandwidth spreading by phase shift keying and differential coherent detection. Short spreading sequences (5 chips/bit) are used to combat multipath induced intersymbol interference. The differential coherent detector is a key element of the modem since it performs IF matched filtering and differentially coherent demodulation of the received signal. An important advantage of this technique is that it reduces cross-correlation noise when short spreading sequences are used. For spreading sequences of length 5 chips/bit, it was observed that the signal obtained at the output of the differential coherent detector contained smaller sidelobes when compared against the signals present at the output of each matched filter.

A commutation signaling differential coherent detector has been implemented as a multi-chip module. This MCM incorporates SAW, analog silicon and thin-film substrate technologies. SAW technology is used to realize IF signal processing functions, specifically, correlation and delay. The SAW devices have been fabricated on 128°-rotated, Y-cut, X-propagating lithium niobate using approximately 1 micron line widths. Analog silicon (i.e., RF IC) technology is used to implement the high-speed bilinear multipliers needed for differential coherent detection as well as the low-impedance buffers used to drive the multipliers. A thin-film alumina substrate is used as the carrier for mounting the SAW and RF IC dies and other surface-mount components that comprise the differential coherent detector. An impedance matching network has been integrated on the substrate. This matching network employs quarter-wave matching transformers and 50 ohm transmission line stubs. These have been defined as patterned coplanar waveguide on the thin-film substrate.

Phase-coded SAW devices are used in the differential coherent detector for matched filtering and delay. Two IDT structures were investigated: (1) a structure that uses wideband tapped output IDTs (i.e., a wideband structure), and (2) a structure that uses narrowband coded output IDTs (i.e., a narrowband structure). Due to photolithography limitations, the IDTs are of the single-electrode type. It was observed experimentally that the measured impulse response of the SAW devices exhibited two main types of distortion: (1) a definite taper (amplitude "droop") over the duration of the impulse response, and (2) significant trailing energy after the main pulse. It was concluded that acoustic reflections at the metal electrode edges (i.e., finger reflections) are the major cause of these distortions. As previously explained, the devices that employed a wideband structure had the best overall performance. An important effect of finger reflections is to reduce the peak-to-maximum-sidelobe ratio of the matched filter autocorrelation response.

In order to experimentally investigate the effect of different metal systems on SAW filter performance, devices were fabricated using 400Å and 1000Å thick aluminum film and 400Å and 1000Å thick gold film on 128° YX-lithium niobate. The performance of the 1000Å gold device was poor. Its center frequency was around 868 MHz. The performance of the 400Å gold device was similar to the performance of the 400Å and 1000Å aluminum devices. The center frequencies of the 400Å aluminum device, 1000Å aluminum device and 400Å gold device were around 996 MHz, 991 MHz and 975 MHz respectively. Wideband devices were chosen to implement the SAW matched filters in the differential coherent detector. The matched filter performance was satisfactory to demonstrate the practical feasibility of a SAW-based differential coherent detector operating at a bit rate of 40 Mb/s, a chip rate of 200 Mc/s and an IF frequency of 1 GHz.

Part II

Performance Analysis of Multicarrier Commutation Signaling Over a Nonlinear Multipath Channel

Chapter 8

Introduction - Part II

Commutation signaling [7], [8], [10], [11], [12] is a bandwidth expanding modulation scheme that is robust to multipath induced intersymbol interference. By using multipath diversity combining, commutation signaling exploits the time diversity that is inherent in a multipath propagation medium. The specific commutation signaling scheme considered in this thesis uses binary differential phase shift keying (DPSK), differential coherent detection, and RAKE multipath diversity combining.

Part I of this thesis has demonstrated the practical feasibility of implementing a commutation signaling differential coherent detector using SAW and analog silicon technologies. The following performance targets were achieved: (1) a bit rate of 40 Mb/s, (2) a chip rate of 200 Mc/s, and (3) an operating frequency (IF frequency) of 1 GHz. A multicarrier commutation signaling modem uses frequency-division multiplexed subcarriers to combine a number of lower rate bit streams to achieve high-rate data transmission. The frequency bands corresponding to individual subcarriers may or may not overlap. In order to achieve an aggregate bit rate of 160 Mb/s, it is necessary to use four subcarriers. Whenever multicarrier modulation is used, the effect of intermodulation distortion (IMD) needs to be considered. In most radio systems, the primary source of intermodulation distortion is the power amplifier. In order to reduce IMD to an acceptable level, the power amplifier needs to be backed off. The amount of backoff required determines the cost and complexity of the transmitter.

Part II of this thesis describes the results of a computer simulation study of the performance of multicarrier commutation signaling over a nonlinear multipath channel. Results that show the effect of subcarrier spacing and intermodulation distortion on modem performance are presented by means

of eye diagrams and error rate curves. The computer simulation study was performed using the Signal Processing WorkSystem® (SPW™) software simulation package of the Alta Group™ (formerly Comdisco Systems, Inc.) of Cadence Design Systems, Inc. The nonlinear multipath channel is modeled as the cascade combination of a transmit filter, a memoryless envelope nonlinearity, a Rummier multipath channel block, and a receive filter. Additive channel noise is assumed to be Gaussian and white. The simulations were performed using lowpass equivalent signals and systems (i.e., complex envelope techniques).

Part II of this thesis is organized in the following way. Chapter 9 describes the simulation model. Chapter 10 compares simulation results for a single carrier commutation signaling modem against those for a four-subcarrier commutation signaling modem that uses Nyquist channel filtering with an excess bandwidth of 50% (i.e., $BW = 1.5R_c$, where BW is the channel bandwidth and R_c is the chip rate). The single carrier commutation signaling modem does not use any channel filtering. Chapter 11 presents performance results for a four-subcarrier commutation signaling modem that uses Nyquist channel filtering over a *linear* multipath channel. Simulations were carried out for a subcarrier spacing corresponding to 50%, 20% and 0% excess bandwidth (i.e., channel bandwidths of $1.5R_c$, $1.2R_c$ and R_c respectively). Chapter 12 presents performance results for a four-subcarrier commutation signaling modem that uses Nyquist channel filtering over a *nonlinear* multipath channel that introduces intermodulation distortion. Chapter 13 summarizes the work done for Part II. Chapter 14 presents the overall conclusions for Parts I and II. A bibliography is included following Chapter 14. Appendix A presents the SPW™ signal flow block diagrams representing the simulation model of the multicarrier commutation signaling modem. The parameter values shown in the block diagrams are default values. These values are adjusted in accordance with the requirements of the overall simulation model.

Chapter 9

Simulation Model

This chapter describes the simulation model used to study the performance of multicarrier commutation signaling. The simulation model was developed using the SPW Π software simulation package of the Alta Group Π of Cadence Design Systems, Inc. The specific commutation signaling scheme that was studied uses binary differential phase shift keying, differential coherent detection and RAKE multipath diversity combining. The nonlinear multipath channel is modeled as the cascade combination of a transmit filter, a memoryless envelope nonlinearity, a Rummler multipath channel block, and a receive filter. Additive channel noise is assumed to be Gaussian and white. The simulations were performed using lowpass equivalent signals and systems (i.e., complex envelope techniques).

9.1 System Description

The four-subcarrier commutation signaling transmitter produces a passband spectrum with a carrier frequency f_c . A passband signal can be represented by a complex analytic signal of the form

$$g_a(t) = \tilde{g}(t)e^{j2\pi f_c t} = [g_c(t) + jg_s(t)]e^{j2\pi f_c t} \quad (9.1)$$

for which the corresponding real signal is

$$g(t) = \text{Re}[g_a(t)] = g_c(t)\cos(2\pi f_c t) - g_s(t)\sin(2\pi f_c t) \quad (9.2)$$

where $g_c(t)$ and $g_s(t)$ are real baseband signals. Therefore, the passband signal $g(t)$ can be represented by its in-phase and quadrature baseband signals $g_c(t)$ and $g_s(t)$, or equivalently, by its complex baseband signal (i.e., complex envelope) $\tilde{g}(t)$.

For the four-subcarrier commutation signaling simulation model, the complex envelope consists of four complex subcarriers separated by a frequency spacing Δf . The differentially encoded information sequence is first combined with the commutation signaling spreading sequences. The combined sequences are then used to phase modulate the complex subcarrier at $f_0 = 0$. The subcarrier data rate is normalized to 1.0 Hz (bps) and the chip rate to 5.0 Hz (bps). Thus, the bandwidth expansion factor or processing gain is 5. The other complex subcarriers at frequencies $f_0 - \Delta f$, $f_0 + \Delta f$ and $f_0 + 2\Delta f$ are phase modulated with random binary data at the chip rate.

During the i -th signaling interval, the complex envelope of the transmitted signal can be written as

$$\tilde{s}_i(t) = \alpha_i(t)e^{j2\pi(f_0 - \Delta f)t} + b_i u_i(t)e^{j2\pi f_0 t} + \beta_i(t)e^{j2\pi(f_0 + \Delta f)t} + \gamma_i(t)e^{j2\pi(f_0 + 2\Delta f)t} \quad (9.3)$$

where $\alpha_i(t)$, $\beta_i(t)$ and $\gamma_i(t)$ each consist of 5 chips of duration T_c and of amplitude ± 1 . The amplitude of each chip is randomly chosen by a toss of a fair coin. Each successive signaling interval, $u_i(t)$ is cyclically chosen from the commutation signaling waveform set $\{u_0(t), u_1(t), u_2(t)\}$. b_i represents the i -th differentially encoded data bit. Fig. 9.1 shows the composite complex baseband spectrum for both $\Delta f = 7.5$ Hz and $\Delta f = 5.0$ Hz.

Fig. 9.2 depicts the simulation block diagram for the four-subcarrier commutation signaling system. The commutation signaling DPSK modulator

block differentially encodes the information bit stream and generates a binary, phase shift keyed, direct sequence spread spectrum signal. The spreading sequences correspond to one of three cyclically repeating commutation signaling codes. The resulting spread spectrum signal is filtered by the transmit filter which provides pulse spectral shaping. The transmit filter is a square root raised cosine filter with rolloff factor equal to 0.5. The commutation DPSK modulated/filtered signal is combined with the other DPSK modulated/filtered subcarriers to generate the composite complex baseband signal.

Figure 9.1. Spectrum of complex composite baseband signal.

Figure 9.2. Four-subcarrier commutation signaling system simulation block diagram.

The composite complex baseband signal is processed by a memoryless envelope nonlinearity. The effect of the nonlinearity is to create both in-band and out-of-band intermodulation distortion in the output spectrum. The output signal from the nonlinear block is input to the multipath channel block. The multipath channel block simulates, at complex envelope level, a two-path, frequency-selective, nonfading multipath channel. Additive channel white Gaussian noise is simulated using an equivalent complex baseband noise model. At the receive end of the equivalent baseband complex transmission system, the complex received signal is filtered by a square root raised cosine filter which is matched to the transmit filter. The output of the receive filter is processed by the Generalized Likelihood (GL) receiver for differentially encoded binary commutation signaling [12]. The received data from the output of the GL receiver is compared against the transmitted data by the bit error rate block, and an error count is generated.

9.2 Commutation Signaling DPSK Modulator

Figure 9.3. Commutation signaling DPSK modulator.

Fig. 9.3 shows the block diagram of the commutation signaling DPSK modulator. The information bit stream is first differentially encoded so that phase transitions between successive encoded data bits represent the original data sequence. Each encoded data bit is then modulated by a commutation signaling codeword. The commutation signaling codewords are bandwidth spreading sequences chosen from a codebook of size three and length five. The following codewords were used in the simulations:

[10110, 11110, 10001]

9.3 Commutation Signaling DPSK Demodulator

Figure 9.4. Commutation signaling DPSK demodulator.

Fig. 9.4 shows the block diagram for the commutation signaling DPSK demodulator. The structure for the demodulator is derived from an application of the Generalized Likelihood principle [12]. For differentially encoded binary commutation signaling that uses N commutation waveforms, the optimal GL receiver consists of N matched filters followed by N square-law equal weight RAKE combiners. For the simulations carried out in this report, it was assumed that the multipath channel delay was known exactly. As a consequence, for $N=3$, the GL receiver takes the simplified form shown in Fig. 9.4.

In Fig. 9.4, each filter is matched to one commutation signaling waveform. For each matched filter pair, differentially coherent demodulation of the received signal is achieved by multiplying the conjugated output of one matched filter by the delayed (by symbol period T_s) output of the other matched filter. The output signal from the complex multiplier is processed by a block that extracts the real part. Due to the bandwidth spreading nature of the commutation waveforms, the output signal from the real part selector block contains narrow autocorrelation signal peaks. The number of peaks

corresponds to the number of paths in the multipath channel (assumed to be two). These peaks are separated in time by the multipath channel delay.

The purpose of the RAKE combiner is to sum these autocorrelation signal peaks. The RAKE combiner has an impulse response which is matched to the multipath channel impulse response. The output of the RAKE receiver is sampled at the instant when the autocorrelation peaks due to both the main path and the delayed path are aligned. The sampled signal is compared against a fixed threshold, and a decision is made as to whether a binary 0 or 1 was transmitted. The demodulated data from all the RAKE combiners is then multiplexed into a serial bit stream.

9.4 Multipath Channel Block

The multipath channel block models a two-path, nonfading multipath channel with the following complex, lowpass impulse response

$$\tilde{h}(t) = \alpha_0 e^{j\theta_0} \delta(t - \tau_0) + \alpha_1 e^{j\theta_1} \delta(t - \tau_1) \quad (9.4)$$

where α_i is the amplitude of the i th path, θ_i is the associated phase shift, and τ_i is the path time delay. The channel output response $\tilde{r}(t)$ to a complex input signal $\tilde{s}(t)$ is given by the time convolution of $\tilde{s}(t)$ and $\tilde{h}(t)$

$$\tilde{r}(t) = \tilde{s}(t) * \tilde{h}(t) = \alpha_0 e^{j\theta_0} \tilde{s}(t - \tau_0) + \alpha_1 e^{j\theta_1} \tilde{s}(t - \tau_1) \quad (9.5)$$

The two-path multipath simulation model is based on the Rummier model which has a 180° phase shift in the delayed path. Fig. 9.5 shows the SPWII simulation model for a two-path multipath channel where the path power difference is 3 dB, the phase difference is 180°, and the relative time delay is $1.6T_S$ (T_S is the symbol period). The time delay of the delayed path is specified in the complex variable delay block.

Figure 9.5. Two-path multipath simulation model at complex envelope level.

9.5 Memoryless Bandpass Nonlinearity

Figure 9.6. Simulation model of envelope nonlinearity.

A memoryless bandpass nonlinearity exhibits both nonlinear gain (AM/AM) and amplitude-to-phase conversion (AM/PM). Let the input to the nonlinearity be a passband signal

$$x(t) = A(t) \cos[2\pi f_c t + \phi(t)] \quad (9.6)$$

Then, the output of the nonlinearity can be expressed as [25]

$$y(t) = f(A(t)) \cos[2\pi f_c t + \phi(t) + g(A(t))] \quad (9.7)$$

where $f(A(t))$ is the nonlinear gain and $g(A(t))$ is the AM/PM conversion.

The complex envelope of the input signal is given by

$$\tilde{x}(t) = A(t) e^{j\phi(t)} \quad (9.8)$$

and the complex envelope of the output signal has the form

$$\tilde{y}(t) = f(A(t)) e^{j[\phi(t) + g(A(t))]} \quad (9.9)$$

The SPWII simulation block diagram that implements the envelope (bandpass) nonlinearity is shown in Fig. 9.6. The input to the block is $\tilde{x}(t)$ and the output from the block is $\tilde{y}(t)$. In an envelope nonlinearity, the nonlinear

part of the output depends only on the envelope $A(t)$ of the input signal. Both the AM/AM conversion $f(A(t))$ and the AM/PM conversion $g(A(t))$ can be approximated by third order polynomials as follows

$$\begin{aligned} f(A(t)) &= c_1 A(t) + c_2 (A(t))^2 + c_3 (A(t))^3 \\ g(A(t)) &= k_1 A(t) + k_2 (A(t))^2 + k_3 (A(t))^3 \end{aligned} \quad (9.10)$$

For the simulations carried out in this study, the AM/PM effect was not considered (i.e., $k_1 = k_2 = k_3 = 0$). The amplitude transfer characteristic was modeled by the following polynomial approximation

$$c_1 = 1; \quad c_2 = 0; \quad c_3 = -0.031$$

For a single complex tone input, this amplitude transfer characteristic gives rise to a power transfer curve that increases nonlinearly until a point is reached at which any further increase in input power results in a decreasing output power. This point of maximum output power is referred to as saturation. The 1 dB compression point refers to the power level at which the actual output power drops 1 dB below the maximum power. The 3 dB compression point is defined in an analogous way. The operating point on the power transfer curve can be specified in terms of input or output backoff from saturation. For example, 3 dB output backoff means that only half of the possible output power is used.

Let the input signal to the nonlinear block consist of a single complex test tone

$$\begin{aligned} \tilde{x}(t) &= A(t)e^{j\phi(t)} \\ &= Ke^{j\omega_1 t} \end{aligned} \quad (9.11)$$

Then the output signal is given by

$$\begin{aligned} \tilde{y}(t) &= f(A(t))e^{j[\phi(t)+g(A(t))]} \\ &= [c_1 A(t) + c_3 (A(t))^3]e^{j\phi(t)} \\ &= [c_1 K + c_3 K^3]e^{j\omega_1 t} \end{aligned} \quad (9.12)$$

Figure 9.7. Power transfer characteristic of envelope nonlinearity.

The average power of the complex input signal is $\bar{P}_{in} = |K|^2$ or $20 \log_{10}|K|$ in dB. The average power of the complex output signal is $\bar{P}_{out} = |c_1 K + c_3 K^3|^2$ or $20 \log_{10}|c_1 K + c_3 K^3|$ in dB. Fig. 9.7 shows the resulting power transfer characteristic when $c_1 = 1$, $c_2 = 0$ and $c_3 = -0.031$.

Now consider the case when two equal complex test tones at frequencies f_1 and f_2 are applied to the input of the nonlinear block. Due to the cubic nonlinearity, the output signal will contain third-order intermodulation distortion products at frequencies $2f_1 - f_2$ and $2f_2 - f_1$.

The input signal is given by

$$\begin{aligned} \tilde{x}(t) &= K e^{j\omega_1 t} + K e^{j\omega_2 t} \\ &= K(\cos \omega_1 t + \cos \omega_2 t) + jK(\sin \omega_1 t + \sin \omega_2 t) \\ &= x_c(t) + jx_s(t) \end{aligned} \quad (9.13)$$

In polar form, the input signal can be represented as

$$\tilde{x}(t) = r(t) e^{j\theta(t)} \quad (9.14)$$

where

$$\begin{aligned} r(t) &= |\tilde{x}(t)| = \sqrt{x_c^2(t) + x_s^2(t)} \\ \theta(t) &= \arg \tilde{x}(t) = \tan^{-1} \frac{x_s(t)}{x_c(t)} \end{aligned} \quad (9.15)$$

The modulus $r(t)$ and phase $\theta(t)$ of $\tilde{x}(t)$ are obtained as follows

$$\begin{aligned} r(t) &= \sqrt{K^2(\cos \omega_1 t + \cos \omega_2 t)^2 + K^2(\sin \omega_1 t + \sin \omega_2 t)^2} \\ &= |K| \sqrt{2(1 + \cos \omega_1 t \cos \omega_2 t + \sin \omega_1 t \sin \omega_2 t)} \\ &= 2|K| \sqrt{\frac{1 + \cos[(\omega_1 - \omega_2)t]}{2}} \\ &= 2 \left| K \cos \left[\frac{1}{2}(\omega_1 - \omega_2)t \right] \right| \\ \theta(t) &= \arg \tilde{x}(t) = \tan^{-1} \left[\frac{x_s(t)}{x_c(t)} \right] \\ &= \tan^{-1} \left[\frac{K(\sin \omega_1 t + \sin \omega_2 t)}{K(\cos \omega_1 t + \cos \omega_2 t)} \right] \\ &= \tan^{-1} \left\{ \frac{2 \sin \left[\frac{1}{2}(\omega_1 + \omega_2)t \right] \cos \left[\frac{1}{2}(\omega_1 - \omega_2)t \right]}{2 \cos \left[\frac{1}{2}(\omega_1 + \omega_2)t \right] \cos \left[\frac{1}{2}(\omega_1 - \omega_2)t \right]} \right\} \\ &= \tan^{-1} \left\{ \tan \left[\frac{1}{2}(\omega_1 + \omega_2)t \right] \right\} \\ &= \frac{1}{2}(\omega_1 + \omega_2)t + k\pi \end{aligned} \quad (9.16)$$

where k is defined by the signs of $x_c(t)$ and $x_s(t)$ as follows

$x_c(t)$	$x_s(t)$	k
+	+	0
+	-	0
-	+	1
-	-	1

Therefore, the complex input signal can be expressed as

$$\begin{aligned}\tilde{x}(t) &= r(t)e^{j\theta(t)} \\ &= 2\left|K \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right]\right| e^{j\left[\frac{1}{2}(\omega_1 + \omega_2)t + k\pi\right]}\end{aligned}\quad (9.17)$$

It should be noted that $\theta(t)$ jumps discontinuously by the amount $\pm\pi$ whenever $K \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right]$ goes through zero. Thus, $\tilde{x}(t)$ can also be written as

$$\begin{aligned}\tilde{x}(t) &= A(t)e^{j\phi(t)} \\ &= 2K \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right] e^{j\frac{1}{2}(\omega_1 + \omega_2)t}\end{aligned}\quad (9.18)$$

For $f(A(t)) = c_1 A(t) + c_3 (A(t))^3$ and $g(A(t)) = 0$, the complex output signal is obtained from the complex input signal as follows

$$\begin{aligned}\tilde{y}(t) &= f(A(t))e^{j[\phi(t) + g(A(t))]} \\ &= [c_1 A(t) + c_3 (A(t))^3] e^{j\phi(t)} \\ &= \left\{2c_1 K \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right] + 8c_3 K^3 \cos^3\left[\frac{1}{2}(\omega_1 - \omega_2)t\right]\right\} e^{j\left[\frac{1}{2}(\omega_1 + \omega_2)t\right]} \\ &= \left\{2c_1 K \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right] + 6c_3 K^3 \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right] + 2c_3 K^3 \cos\left[\frac{3}{2}(\omega_1 - \omega_2)t\right]\right\} \cdot \\ &\quad e^{j\left[\frac{1}{2}(\omega_1 + \omega_2)t\right]} \\ &= \left\{(2c_1 K + 6c_3 K^3) \cos\left[\frac{1}{2}(\omega_1 - \omega_2)t\right] + 2c_3 K^3 \cos\left[\frac{3}{2}(\omega_1 - \omega_2)t\right]\right\} e^{j\left[\frac{1}{2}(\omega_1 + \omega_2)t\right]} \\ &= (c_1 K + 3c_3 K^3) \left\{e^{j\left[\frac{1}{2}(\omega_1 - \omega_2)t\right]} + e^{-j\left[\frac{1}{2}(\omega_1 - \omega_2)t\right]}\right\} e^{j\left[\frac{1}{2}(\omega_1 + \omega_2)t\right]} \\ &\quad + c_3 K^3 \left\{e^{j\left[\frac{3}{2}(\omega_1 - \omega_2)t\right]} + e^{-j\left[\frac{3}{2}(\omega_1 - \omega_2)t\right]}\right\} e^{j\left[\frac{1}{2}(\omega_1 + \omega_2)t\right]} \\ &= (c_1 K + 3c_3 K^3) (e^{j\omega_1 t} + e^{j\omega_2 t}) + c_3 K^3 \{e^{j[(2\omega_1 - \omega_2)t]} + e^{j[(2\omega_2 - \omega_1)t]}\}\end{aligned}\quad (9.19)$$

From the above, it is clear that, if K is increased sufficiently, the intermodulation distortion (IMD) will become significant since the desired output (at f_1 or f_2) varies linearly with K , while the third-order intermodulation

distortion output (at $2f_1 - f_2$ or $2f_2 - f_1$) varies as K^3 . It is also seen that, when c_1 and c_3 have opposite signs, the desired output (at f_1 or f_2) will saturate when K is made sufficiently large.

The average power in the desired linearly amplified output (at either f_1 or f_2) is $\bar{P}_{lin} = |c_1 K|^2$ or $20 \log_{10} |c_1 K|$ in dB. The average power in each of the third-order IMD products is $\bar{P}_{imd} = |c_3 K^3|^2$ or $20 \log_{10} |c_3 K^3|$ in dB. Hence, the third-order intermodulation product power increases three times as quickly as the desired output power. For $c_1 = 1$, $c_2 = 0$ and $c_3 = -0.031$, the average power in each of the third-order IMD products is 30 dB below the desired output power (at f_1 or f_2) when the input power (at f_1 or f_2) is 0 dB.

With reference to Fig. 9.7, the saturation point of the power transfer characteristic occurs at an output power of approximately 6.5 dB. An input power level of 3 dB corresponds to an output power level of approximately 2 dB. Hence, this operating point corresponds to an output backoff of about 4.5 dB.

Fig. 9.8 shows the output spectrum of the envelope nonlinearity for a baseband input signal consisting of four modulated subcarriers (7.5 Hz and 5.0 Hz subcarrier spacing). The nonlinearity introduces both in-band and out-of-band intermodulation interference. The input average power level, input peak-to-average power ratio, output power level, output peak-to-average power ratio and gain compression are summarized in Table 9.1.

Table 9.1. Nonlinearity input and output power levels.

	7.5 Hz subcarrier spacing	5.0 Hz subcarrier spacing
Average input power	3 dB	3 dB
Input peak-to-average power ratio	8.6 dB	8.7 dB
Average output power	1.96 dB	1.96 dB
Output peak-to-average power ratio	4.8 dB	4.8 dB
Gain compression	1.04 dB	1.04 dB

Figure 9.8. Output spectrum of envelope nonlinearity.

Chapter 10

Simulation Results - Single Carrier vs. Four-Subcarrier Modem

This chapter presents a comparison of simulation results for a single carrier commutation signaling modem vs. a four-subcarrier modem that uses Nyquist channel filtering with an excess bandwidth of 50% (i.e., $BW = 1.5R_c$, where BW is the channel bandwidth and R_c is the chip rate). The single carrier commutation signaling modem does not use any channel filtering (i.e., the transmitted signal has a constant envelope). For the four-subcarrier commutation signaling modem, the complex composite baseband signal consists of four phase modulated complex subcarriers that are equally spaced at 7.5 Hz. The composite baseband signal has a peak-to-average power ratio of 8.2 dB. The overall bandwidth of the composite baseband signal is 30.0 Hz (4 X 7.5 Hz). For a 7.5 Hz subcarrier spacing, the individual subcarrier spectra do not overlap. As a consequence, there is no crosstalk between message signals in adjacent channels.

The commutation signaling receiver uses ideal RAKE combining (i.e., the multipath channel delay is known exactly). For simulation purposes, the two-path multipath channel was characterized by the parameters shown in Table 10.1. Both the direct path and delayed path have equal signal strength. The delayed path introduces a 270° phase shift relative to the direct path. The multipath channel delay is $1.6T_S$ (T_S is the symbol time).

Table 10.1. Two-path multipath parameters.

	α_i	θ_i	τ_i
Direct Path ($i = 0$)	1	0°	0
Delayed Path ($i = 1$)	1	270°	$1.6T_S$

Figure 10.1. Error rate performance for binary DPSK commutation signaling over a single path and a two-path multipath nonfading channel (single carrier; no channel filtering).

Figure 10.2. Error rate performance for binary DPSK commutation signaling over a single path and a two-path multipath nonfading channel (four subcarriers - 7.5 Hz subcarrier spacing; Nyquist channel filtering).

Fig. 10.1 shows the error rate performance, for the single carrier commutation signaling modem, over both a single path nonfading channel and a two-path multipath nonfading channel. For the single path nonfading channel, the error rate performance of binary DPSK commutation signaling is identical to that of binary DPSK (solid curve in Fig. 10.1). For the two-path, frequency-selective, nonfading multipath channel described above, the error rate performance (dashed curve) is approximately 1.5-2 dB worse than theoretical DPSK. The loss in performance results from intersymbol interference caused by the multipath. The probability of bit error for each simulation point (i.e., SNR value) was obtained by averaging the simulation results over ten trials. The number of transmitted symbols per trial varied from $40-160/P_e$, where P_e represents the average probability of bit error. The standard deviation of the average probability of bit error about the mean is shown by the error bars in Fig. 10.1.

Fig. 10.2 shows the error rate performance, for the four-subcarrier commutation signaling modem, over both a single path nonfading channel and a two-path multipath nonfading channel characterized by the same parameters as shown in Table 10.1. Comparison of Fig. 10.1 and Fig. 10.2 shows that the performance of the four-subcarrier commutation signaling modem with 7.5 Hz subcarrier spacing is essentially identical to that of the single carrier commutation signaling modem. This is expected since, for a 7.5 Hz subcarrier spacing, individual subcarrier spectra do not overlap.

Fig.'s 10.3-10.8 present eye diagrams at the output of the RAKE receiver for the four-subcarrier commutation signaling modem. Fig. 10.3 shows the eye diagram for a single path, noiseless, nonfading channel. At the sampling time, the eye is almost completely open. There is a negligible amount of intersymbol interference due to the finite length of the square root raised cosine FIR filters in the transmitter and the receiver. Hence, the degradation in performance is expected to be nil. This is confirmed by the error rate performance results in Fig. 10.2.

Fig.'s 10.4-10.7 present eye diagrams at the output of the RAKE receiver for a two-path, noiseless, nonfading multipath channel, where the path power difference is 3 dB and the relative phase differences between the delayed path and the direct path are 0° , 90° , 180° and 270° respectively. The

time delay of the delayed path relative to the direct path is $1.6T_s$. When the relative phase difference is 0° or 180° (Fig.'s 10.4 and 10.6), the signal from the delayed path is "in phase" with the signal from the direct path. This represents the worst case condition for intersymbol interference. When the relative phase difference is 90° or 270° (Fig.'s 10.5 and 10.7), the signal from the delayed path is "in quadrature" with the signal from the direct path. This represents the best case condition for intersymbol interference.

Fig. 10.8 presents the eye diagram at the output of the RAKE receiver for a single path, noiseless, nonlinear, nonfading channel. The nonlinearity introduces third-order intermodulation distortion. It is characterized by $c_1 = 1$, $c_2 = 0$, $c_3 = -0.031$ (see Fig. 9.7). Comparison of Fig. 10.8 with Fig. 10.3 shows that there is some intersymbol interference due to the intermodulation distortion.

Figure 10.3. Eye diagram at the RAKE receiver output for a single path nonfading channel (7.5 Hz subcarrier spacing).

Figure 10.4. Eye diagram at the RAKE receiver output for a two-path multipath nonfading channel (7.5 Hz subcarrier spacing; 0° phase difference).

Figure 10.5. Eye diagram at the RAKE receiver output for a two-path multipath nonfading channel (7.5 Hz subcarrier spacing; 90° phase difference).

Figure 10.6. Eye diagram at the RAKE receiver output for a two-path multipath nonfading channel (7.5 Hz subcarrier spacing; 180° phase difference).

Figure 10.7. Eye diagram at the RAKE receiver output for a two-path multipath nonfading channel (7.5 Hz subcarrier spacing; 270° phase difference).

Figure 10.8. Eye diagram at the RAKE receiver output for a single path, nonlinear, nonfading channel (7.5 Hz subcarrier spacing).

Chapter 11

Simulation Results - Four-Subcarrier Modem (Without Intermodulation Distortion)

This chapter presents simulation results for a four-subcarrier commutation signaling modem that uses Nyquist channel filtering over a linear multipath channel. The composite baseband signal consists of four, equally spaced, phase modulated complex subcarriers. The subcarrier frequency separation is Δf . Simulations were carried out for $\Delta f = 7.5$ Hz, $\Delta f = 6.0$ Hz and $\Delta f = 5.0$ Hz. For $\Delta f = 7.5$ Hz (50% excess bandwidth), the overall bandwidth of

Figure 11.1. Error rate performance for binary DPSK commutation signaling over a two-path multipath, nonfading, linear channel (four subcarriers; Nyquist channel filtering).

the composite baseband signal is 30.0 Hz. The individual subcarrier spectra do not overlap. Therefore, there is no crosstalk between adjacent channels. For $\Delta f = 6.0$ Hz (20% excess bandwidth), the overall bandwidth of the composite baseband signal is 24.0 Hz. The individual subcarrier spectra overlap to some extent. As a consequence, there is some crosstalk between message signals in adjacent channels. The bandwidth reduction is 20% when compared against the case when the subcarrier spacing is 7.5 Hz. For $\Delta f = 5.0$ Hz (0% excess bandwidth), the overall bandwidth of the composite baseband signal is 20.0 Hz. The individual subcarrier spectra overlap to a significant extent. This results in significant adjacent channel interference. The bandwidth reduction is 33% when compared against the case when the subcarrier spacing is 7.5 Hz. The simulations for $\Delta f = 5.0$ Hz were performed with adjacent subcarriers offset in phase by 90° .

Fig. 11.1 shows the error rate performance, for the four-subcarrier commutation signaling modem, over a linear, two-path multipath, nonfading channel. For simulation purposes, the two-path multipath channel was characterized by the parameters shown in Table 11.1. The path power difference between the direct path and the delayed path is 3 dB. There is a 0° relative phase difference between the direct path and the delayed path. This corresponds to the worst case condition which results in the greatest amount of intersymbol interference. The multipath channel delay is $1.6T_s$.

Table 11.1. Two-path multipath parameters.

	α_i	θ_i	τ_i
Direct Path ($i = 0$)	1	0°	0
Delayed Path ($i = 1$)	.707	0°	$1.6T_s$

Fig. 11.1 shows that the error rate performance for $\Delta f = 6.0$ Hz is only marginally worse than for $\Delta f = 7.5$ Hz (approx. 1 dB worse for a bit error rate of 10^{-4}). The performance is, however, significantly worse for $\Delta f = 5.0$ Hz. In order to achieve a bit error rate of 10^{-4} , for example, the signal-to-noise ratio needs to be increased by approximately 6 dB (compared to $\Delta f = 7.5$ Hz). As the subcarrier frequency spacing is decreased, adjacent channel crosstalk increases. This results in an irreducible noise floor due to intersymbol interference. Any

additional channel noise simply adds to this noise floor. At low signal-to-noise ratios, error rate performance is dominated by channel noise. At high signal-to-noise ratios, error rate performance is dominated by adjacent channel intersymbol interference.

Fig. 11.2 presents eye diagrams at the RAKE receiver output for a linear, single path, nonfading channel for subcarrier spacings of 5.0, 6.0 and 7.5 Hz. As the subcarrier spacing is decreased, the eye becomes more closed due to adjacent channel intersymbol interference. Fig. 11.3 presents the same eye diagrams when the subcarriers in the two adjacent channels are phase offset by 90° . For significantly overlapped subcarrier spectra (e.g., 5.0 Hz subcarrier spacing), it is seen that the use of carrier phase offsets in the adjacent channels will increase the size of the eye opening and, hence, will have a beneficial impact on system error rate performance. This is an interesting result and could be a topic for further study.

Figure 11.2. Eye diagrams at the RAKE receiver output for a linear, single path, nonfading channel (5.0, 6.0 and 7.5 Hz subcarrier spacings; 0° carrier phase offsets).

Figure 11.3. Eye diagrams at the RAKE receiver output for a linear, single path, nonfading channel (5.0, 6.0 and 7.5 Hz subcarrier spacings; 90° carrier phase offsets).

Chapter 12

Simulation Results - Four-Subcarrier Modem (With Inter-modulation Distortion)

This chapter presents simulation results for a four-subcarrier commutation signaling modem that uses Nyquist channel filtering over a nonlinear multipath channel. The composite baseband signal consists of four, equally spaced, phase modulated complex subcarriers. The subcarrier frequency separation is Δf . Simulations were carried out for $\Delta f = 7.5$ Hz, $\Delta f = 6.0$ Hz and $\Delta f = 5.0$ Hz. The simulations for $\Delta f = 5.0$ Hz were performed with adjacent

Figure 12.1. Error rate performance for binary DPSK commutation signaling over a two-path multipath, nonfading, nonlinear channel (four subcarriers; Nyquist channel filtering).

subcarriers offset in phase by 90° . The multipath channel is characterized by the same parameters as shown in Table 11.1. The channel nonlinearity introduces third-order intermodulation distortion. It is characterized by $c_1 = 1$, $c_2 = 0$, $c_3 = -0.031$ (see Fig. 9.7).

Comparison of Fig.'s 11.1 and 12.1 shows that, for all subcarrier frequency offsets, there is only a fraction of a dB degradation in error rate performance due to the channel intermodulation distortion. This result is discussed further below with reference to eye diagrams. Fig.'s 12.2 and 12.4 present eye diagrams at the output of the RAKE receiver for a noiseless, nonfading, two-path multipath, linear channel, where the path power difference is 3 dB, and the associated channel phase differences are 90° and 0° respectively. The subcarrier frequency spacing is 6.0 Hz. Fig.'s 12.3 and 12.5 show the same eye diagrams when the multipath channel has been distorted by the introduction of nonlinear intermodulation interference. Comparison of Fig.'s 12.2 and 12.3 and Fig.'s 12.4 and 12.5 shows that the eye diagrams are essentially unchanged by the intermodulation interference. Hence, the performance loss caused by the intermodulation interference is expected to be nil. This is confirmed by the error rate performance results in Fig. 12.1. Similar results are obtained for a subcarrier frequency offset of 5.0 Hz as shown in Fig.'s 12.6-12.9.

Figure 12.2. Eye diagram at the RAKE receiver output for a linear, two-path multipath, nonfading channel (6.0 Hz subcarrier spacing; 90° phase difference).

Figure 12.3. Eye diagram at the RAKE receiver output for a nonlinear, two-path multipath, nonfading channel (6.0 Hz subcarrier spacing; 90° phase difference).

Figure 12.4. Eye diagram at the RAKE receiver output for a linear, two-path multipath, nonfading channel (6.0 Hz subcarrier spacing; 0° phase difference).

Figure 12.5. Eye diagram at the RAKE receiver output for a nonlinear, two-path multipath, nonfading channel (6.0 Hz subcarrier spacing; 0° phase difference).

Figure 12.6. Eye diagram at the RAKE receiver output for a linear, two-path multipath, nonfading channel (5.0 Hz subcarrier spacing; 90° phase difference).

Figure 12.7. Eye diagram at the RAKE receiver output for a nonlinear, two-path multipath, nonfading channel (5.0 Hz subcarrier spacing; 90° phase difference).

Figure 12.8. Eye diagram at the RAKE receiver output for a linear, two-path multipath, nonfading channel (5.0 Hz subcarrier spacing; 0° phase difference).

Figure 12.9. Eye diagram at the RAKE receiver output for a nonlinear, two-path multipath, nonfading channel (5.0 Hz subcarrier spacing; 0° phase difference).

Chapter 13

Summary - Part II

Part II of this thesis has described the results of a computer simulation study of the performance of multicarrier commutation signaling over a nonlinear multipath channel. Comparative results that show the effect of subcarrier frequency spacing and intermodulation distortion on modem performance were presented by means of eye diagrams and error rate curves.

The performance, over a linear multipath channel, of a multicarrier commutation signaling modem that uses Nyquist channel filtering with a rolloff factor of 0.5 and a channel $BW = 1.5R_c$ (i.e., no spectral overlap) is essentially identical to that of a single carrier commutation signaling modem. When the subcarrier spectra overlap by 20% (channel $BW = 1.2R_c$), the multicarrier commutation signaling modem suffers a 1 dB performance loss. For a spectral overlap of 33% (channel $BW = R_c$), the loss in performance is 6 dB. The multicarrier commutation signaling modem does, however, have an 8.6 dB peak-to-average power ratio disadvantage (over a linear channel) when compared against the single carrier commutation signaling modem which produces an unfiltered constant envelope signal.

The performance of a multicarrier commutation signaling modem over a nonlinear channel that introduces third-order intermodulation distortion is only marginally degraded (a fraction of a dB) when compared against the linear channel performance. This result was obtained for subcarrier spectral overlaps of 0%, 20% and 33% (i.e., channel bandwidths of $1.5R_c$, $1.2R_c$ and R_c respectively). The channel nonlinearity was modeled as a memoryless envelope nonlinearity. The operating point chosen for the simulations corresponds to a (single carrier power transfer curve) output backoff of 4.5 dB. The nonlinearity had a (multicarrier) gain compression of about 1 dB. The compressed multicarrier signal had a peak-to-average power ratio of 4.8 dB.

Chapter 14

Conclusions

This thesis has described a SAW-based commutation signaling modem for broadband indoor wireless communication. This system employs binary differential encoding, bandwidth spreading by phase shift keying and differential coherent detection. Short spreading sequences (5 chips/bit) are used to combat multipath induced intersymbol interference. The differential coherent detector is a key element of the modem since it performs IF matched filtering and differentially coherent demodulation of the received signal. An important advantage of this technique is that it reduces cross-correlation noise when short spreading sequences are used. Part I of this thesis has demonstrated the practical feasibility of implementing a commutation signaling differential coherent detector using SAW and analog silicon technologies. The following performance targets were achieved: (1) a bit rate of 40 Mb/s, (2) a chip rate of 200 Mc/s, and (3) an operating frequency (IF frequency) of 1 GHz.

The commutation signaling differential coherent detector was implemented as a multi-chip module. This MCM incorporates SAW, analog silicon and thin-film substrate technologies. SAW technology was used to realize the IF signal processing functions of correlation and delay. The SAW devices were fabricated on 128°-rotated, Y-cut, X-propagating lithium niobate using approximately 1 micron line widths. Analog silicon technology, based on Nortel's NT25 high-speed bipolar process, was used to implement the high-speed bilinear multipliers needed for differential coherent detection as well as the low-impedance buffers used to drive the multipliers. A thin-film alumina substrate was used as the carrier for mounting the SAW and RF IC dies and other surface-mount components that comprise the differential coherent detector.

Phase-coded SAW devices were used in the differential coherent detector for matched filtering and delay. Two IDT structures were investigated: (1) a structure that uses wideband tapped output IDTs (i.e., a wideband structure), and (2) a structure that uses narrowband coded output IDTs (i.e., a narrowband structure). Due to photolithography limitations, the IDTs were of the single-electrode type. It was observed experimentally that the measured impulse response of the SAW devices exhibited two main types of distortion: (1) a definite taper (amplitude “droop”) over the duration of the impulse response, and (2) significant trailing energy after the main pulse. It was concluded that acoustic reflections at the metal electrode edges (i.e., finger reflections) are the major cause of these distortions. In order to experimentally investigate the effect of different metal systems on SAW filter performance, devices were fabricated using 400Å and 1000Å thick aluminum film and 400Å and 1000Å thick gold film on 128° YX-lithium niobate. The performance of the 1000Å gold devices was poor. The performance of the 400Å gold devices was similar to the performance of the 400Å and 1000Å aluminum devices. For all metal systems, the devices that employed a wideband structure had the best overall performance.

A multicarrier commutation signaling modem uses frequency-division multiplexed subcarriers to combine a number of lower rate bit streams to achieve high-rate data transmission. The frequency bands corresponding to individual subcarriers may or may not overlap. In order to achieve an aggregate bit rate of 160 Mb/s, it is necessary to use four subcarriers, each supporting a data rate of 40 Mb/s. Whenever multicarrier modulation is used, the effect of intermodulation distortion needs to be considered. Part II of this thesis has described the results of a computer simulation study of the performance of multicarrier commutation signaling over a nonlinear multipath channel. Comparative results that show the effect of subcarrier frequency spacing and intermodulation distortion on modem performance were presented by means of eye diagrams and error rate curves.

As expected, the performance of a multicarrier commutation signaling modem that uses Nyquist channel filtering with no spectral overlap is essentially identical to that of a single carrier commutation signaling modem. When the subcarrier spectra overlap by 20%, the multicarrier commutation

signaling modem suffers a 1 dB performance loss. For a spectral overlap of 33%, the loss in performance is 6 dB. The multicarrier commutation signaling modem does, however, have an 8.6 dB peak-to-average power ratio disadvantage (over a linear channel) when compared against the single carrier commutation signaling modem which produces an unfiltered constant envelope signal. The performance of a multicarrier commutation signaling modem over a nonlinear channel that introduces third-order intermodulation distortion is only marginally degraded (a fraction of a dB) when compared against the linear channel performance. This result was obtained for subcarrier spectral overlaps of 0%, 20% and 33%. The operating point chosen for the simulations corresponds to a (single carrier) output backoff of 4.5 dB. The compressed multicarrier signal had a peak-to-average power ratio of 4.8 dB (vs. 8.6 dB for the linear channel case).

Suggestions for Future Work

This thesis has demonstrated the practical feasibility of implementing a commutation signaling differential coherent detector that uses SAW matched filters. This is an important step towards the goal of developing a broadband indoor wireless demonstration system for millimeter-wave access to an ATM LAN. Although the performance of the SAW filters was satisfactory to show proof-of-concept, the performance needs to be improved. The discrepancies between the first-order theoretical predictions and the measured responses are mainly due to the following second-order effects: (1) finger reflections, (2) electromagnetic feed-through, and (3) triple-transit interference. Suggestions for minimizing the distortion caused by each of these second-order effects are given below. Furthermore, the center frequency and symbol delay need to be fine-tuned by fine-tuning the linewidth and inter-tap spacing. This would eliminate the need for thin-film delay lines and an impedance matching network on the MCM substrate. In order to maintain center frequency and symbol delay tolerances within acceptable limits over ambient temperature variations, a heated SAW package would be required.

Finger Reflections

The major contribution to the distortion observed experimentally for the impulse response of the SAW filters is acoustic mismatch at the electrode discontinuities (i.e., finger reflections). There are two approaches to minimizing finger reflections: (1) use double-electrode fingers, and (2) optimize the metal-film thickness. In order to fabricate double-electrode transducers for operation at 1 GHz, sub-micron lithography is required. This capability should be available at Nortel in the near future. The second approach is to optimize the metal thickness. Experimental confirmation of the possibility of implementing a single-electrode IDT without finger reflections for aluminum metallization on 128° YX-LiNbO₃ is given in [26]. The experimental data shows that a metal thickness of about 1650Å should give a strip reflectivity close to zero for a metallization ratio of 50%. An analysis based on the finite element method indicates that a metal thickness of 950Å⁽¹⁾ should give a strip reflectivity close to zero for a 50% strip metallization ratio [27]. Since 1000Å aluminum devices were fabricated and tested, it is uncertain as to how much additional improvement can be achieved by this technique.

Electromagnetic Feed-Through (EMFT)

The EMFT performance of the SAW devices was poor. This can be improved as follows: (1) use a better package design, (2) separate the input and output IDT ground returns, (3) reduce the die size, and (4) optimize the bond wire arrangement. It would also be beneficial to provide greater separation between matched filters on the same die in order to minimize cross-talk.

Triple-Transit Interference (TTI)

The TTI level (for individual devices measured in a 50Ω system) was high. This can be reduced by terminating the output transducer of the SAW device in a low-impedance load (e.g., similar to the lo-Z input buffer amplifier on the BUFMULT RF IC). It is also necessary to optimize the value of the decoupling capacitor between the SAW device output and the lo-Z buffer input in order to tune out any connecting track inductance.

¹This information was not available at the time the decision was made to fabricate devices using 1000Å thick aluminum metallization.

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Appendix A

SPW™ Simulation Model Signal Flow Block Diagrams

4-CH 1-PATH BER TEST BLOCK PARAMETERS

MAIN PARAMETERS:

Computation signal file name 'george1ibroom1_2001600_200
 Eb/No (dB) 7,5
 Roll off factor 0,5
 Channel 1 carrier frequency (Hz) -7,5
 Channel 2 carrier frequency (Hz) 0,0
 Channel 3 carrier frequency (Hz) 7,5
 Channel 4 carrier frequency (Hz) 15,0

4-CH 2-PATH BER TEST BLOCK PARAMETERS		
MAIN PARAMETERS		
Simulation signal file name	georgiadiscom_628000_240	6.8
Channel 1 transition frequency (Hz)		-5.0
Channel 2 transition frequency (Hz)		5.0
Channel 3 transition frequency (Hz)		15.0
Channel 4 transition frequency (Hz)		18.0
Sampling frequency (Hz)		200.0
Symbol delay (samples)		-40
Symbol rate (symbols/sec.)		5.0
Roll off factor		0.5
First input sample time (sec.)		0.1
Channel 1 carrier phase		1.570796327
Channel 2 carrier phase		1.570796327
Channel 3 carrier phase		0.0
Channel 4 carrier phase		0.0
Channel 1 symbol offset delay (samples)		-20
Channel 3 symbol offset delay (samples)		-20
Channel 4 symbol offset delay (samples)		-10
Offset Symbol Type = (1,0) no = 0.0		0.0

3-CH DPSK TRANSMITTER BLOCK PARAMETERS

MAIN PARAMETERS:

Channel 1 translation frequency (Hz)	0.0
Channel 2 translation frequency (Hz)	0.0
Channel 3 translation frequency (Hz)	0.0
Sampling frequency (Hz)	5.0
Symbol delay (samples)	-5
Symbol rate (symbols/sec.)	1.0
Roll off factor	0.5
First input sample time (sec.'s)	0.0

3-CH DPSK TRANSMITTER BLOCK PARAMETERS

MAIN PARAMETERS:

Channel 1 translation frequency (Hz)	-7.5
Channel 2 translation frequency (Hz)	7.5
Channel 3 translation frequency (Hz)	15.0
Sampling frequency (Hz)	200.0
Symbol delay (samples)	-10
Symbol rate (symbols/sec.)	5.0
Roll off factor	0.5
First input sample time (sec., 's)	0.1
Channel 1 carrier phase	0.0
Channel 2 carrier phase	0.0
Channel 3 carrier phase	0.0
Channel 1 symbol offset delay (samples)	-20
Channel 2 symbol offset delay (samples)	-20
Channel 3 symbol offset delay (samples)	-20
Offset symbol? (yes = 1.0; no = 0.0)	0.0

COMMUTATION/DPSK DEMODULATOR BLOCK PARAMETERS

MAIN PARAMETERS:

Symbol delay (samples)	-200
RAKE delay (samples)	-320
Sampling frequency (Hz)	200.0
Symbol rate (symbols/sec.)	5.0
Roll off factor	0.5

COMPLEX ROOT RAISED COSINE FILTER BLOCK PARAMETERS

MAIN PARAMETERS:

Sampling frequency (Hz)	200.0
Chip rate (chips/sec.)	5.0
Roll off factor	0.5

TRANSVERSAL FILTER BLOCK PARAMETERS

MAIN PARAMETERS:

cl 0)	1.0
cl 1)	0.0
cl 2)	0.0
cl 3)	0.0
cl 4)	0.0
Symbol delay (samples)	-10

COMMUTATION/DPSK DETECTOR BLOCK PARAMETERS

MAIN PARAMETERS:

Symbol delay (samples)

-25

2-PATH RAKE RECEIVER BLOCK PARAMETERS

MAIN PARAMETERS:

RAKE delay (samples)

-40

CLOCK BLOCK PARAMETERS

MAIN PARAMETERS:

Clock period (samples)	75
Clock high period (samples)	-25
Start time	0

IMAGE EVALUATION TEST TARGET (QA-3)

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Fax: 716/288-5989

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