MAPPING UWB SIGNAL PROCESSING ONTO SILICON

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ABSTRACT

In this paper, we study the performance of a transmitted reference (TR) system corrupted by a single tone interferer in a multipath noise-free channel. By employing "frequency wrapping", a new architecture, the quadrature downconversion autocorrelation receiver (QDAR) is developed. Unlike a RAKE receiver, the QDAR does not suffer from timing and template matching problems, and it also circumvents processing at high frequencies, thereby reducing the on-chip circuit complexity and power consumption, and offers an elegant solution to narrowband interference rejection. The figure of merit of auto-correlation receivers can be defined by the accuracy of the the time delay element. In this paper, a *cyclic* delay is designed as required for the autocorrelation function in the QDAR. A Padé approximation is used to derive a rational transfer function of a unit delay filter (i.e., 9^{th} order). Finally the performance of the *cyclic* delay is evaluated through circuit level simulations.

1. INTRODUCTION

In 2002, the Federal Communications Commission (FCC) unanimously agreed in allocating the unlicensed spectrum from 3.1-10.6 GHz for UWB applications. Since then, UWB technology has gained much interest and is viewed as a potential candidate for future wireless short-range data communication. The overwhelming attention that UWB technology has received is primarily due to the promise of a very high data-rate system at low cost and low power consumption, and due to its capability of sharing the bandwidth resources [1].

A particular form of UWB communication is impulse radio [2] (ir-UWB), where very short transient pulses (duration in the order of hundreds of picoseconds) are transmitted rather than a modulated carrier. These pulses occupy a bandwidth of a few gigahertz. The issue of co-existence is a subject of controversy as UWB systems transmitting at low spectral densities overlap with the bands of many other narrowband systems.

Sub-optimal, low complexity transmitted reference (TR) autocorrelation receivers (AcRs) proposed 40 years ago [3], have regained popularity as the synchronization procedure is considerably simplified, and the channel estimation challenge completely avoided. Despite the fact the narrowband interference has been acknowledged as a serious impairment for UWB performance, only limited research has been addressed to evaluate the effects on TR schemes [4, 5, 6].

In this paper, we first show that a single-tone interferer can drastically deteriorate the performance of AcRs. Based on our results, a new receiver architecture (i.e. QDAR), which amalgamates auto-correlation and frequency downconversion, is proposed. Through *frequency wrapping*, not only does it solve the issue of narrowband interference (i.e., by precognitive allocation of the NB interferers in the frequency domain) but also avoids high frequency on-chip processing, thus reducing power consumption. The key element in all AcRs is the delay and to save chip area and power consumption, we propose *cyclic* delay line. This approach allows one to implement a single delay-line with only a small delay. Longer delays are by *re-cycling* the signal. At first, a Padé approximation of a rational transfer function is derived and then a cascade of two 9^{th} order filters is selected to implement this delay.

2. TRANSMITTED REFERENCE SIGNALING SCHEME

2.1. The transmission scheme

In this section we briefly introduce the system model for the transmitted reference scheme. The symbol energy is split into N_d doublets, each one of them consisting of two pulses $p(t)$ of duration T_p delayed in time by D_j, $j = 0 \dots N_d - 1$. Each doublet is repeated every T_d. The set of elements D_j is here referred to as the delay hopping code.

A mathematical representation of the UWB transmitted signal $x(t)$ is given by

$$x(t) = \sum_{n=-\infty}^{\infty} p(t - nT_d) + a_n p(t - nT_d - D_{n \mod N_d}).$$

The first pulse of each doublet is used as the signal template for demodulation of the data conveyed by the second pulse. Indeed, the amplitude of the modulated pulse is given as $a_n = b_{\lfloor n/N_d \rfloor} c_{n \mod N_d}$, where $b_{\lfloor n/N_d \rfloor}$ is the binary data which takes values of the set $\{1, -1\}$ and is repeated for each of the N_d doublets, and $c_j, j = 0, \ldots, N_d - 1$ is the j_{th} element of a chip code with alphabet $\{1, -1\}$ and periodicity equal to N_d. The received UWB signal can be written as

$$y(t) = \sum_{n=-\infty}^{\infty} h(t - nT_d) + a_n h(t - nT_d - D_{n \mod N_d}), \tag{1}$$

where $h(t)$ is the received pulse of duration T_h and includes the transmit pulse $p(t)$ with duration T_p, the propagation channel $c(t)$ of duration T_c, and the receive filter $q(t)$ with duration T_q. Hence, we have $h(t) = p(t) \star c(t) \star q(t)$ and $T_h = T_p + T_c + T_q$. We assume here that $T_d > T_h + \max D_j$, such that interference among doublets is avoided.

The received noise $w(t)$ is modeled as white Gaussian noise with spectral density N_0, denoted as $v(t)$, that is filtered by the receive filter $q(t)$, i.e., $w(t) = v(t) \star q(t)$. The overall received signal can thus be written as

$$r(t) = y(t) + w(t).$$

Performance will be measured as a function of E_b/N_0, where E_b is the bit energy before receiver filtering:

$$E_b = 2 \int (p(t) \star c(t))^2 dt,$$

where the factor 2 is due to the fact that we use 2 pulses to represent a single bit. Note that the measure E_b/N_0 is independent of the receive filter and thus allows us to compare the performance of different receive filters.

In the following sections, we limit ourselves to the first symbol $b_0 = b$, i.e., we assume that $0 \leq t < N_d T_d$. We then describe some detectors to estimate this bit b.

2.2. The conventional autocorrelation principle

The conventional autocorrelation receiver consists of N_d branches, each one provided with a delay-line matched to one of the elements of the delay hopping code $\{D_j\}$ followed by an integrator. The output of the j_{th} receiver branch is described by

$$\begin{aligned} z_j &= \int_{jT_d+D_j}^{jT_d+D_j+T_I} r(t) r(t - D_j) dt \qquad (2) \\ &= a_j \int_0^{T_I} h^2(t) dt + \mu_j, \end{aligned}$$

where μ_j includes all possible self-interference and noise terms. Thus, the unmodulated pulse, which forms some kind of a noise UWB channel estimate, is first time-aligned and then correlated with the modulated pulse. All the outputs z_j are then coherently combined to form the decision variable

$$z = \sum_{j=0}^{N_d-1} c_j z_j = b N_d \int_0^{T_I} h^2(t) dt + \mu,$$

where

$$\mu = \sum_{j=0}^{N_d-1} c_j \mu_j.$$

An estimate of b is then found by

$$\hat{b} = \text{sign}(z).$$

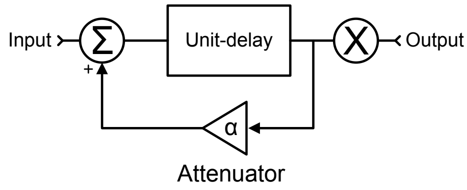

Figure 1: *Cyclic* Delay

2.3. The cyclic autocorrelation principle

The conventional autocorrelation receiver requires N_d branches, which could lead to a highly complex receiver. Moreover, some of the delay-lines have to implement a large delay. Such devices come with a large size and power consumption, and they can not be implemented very accurately. To avoid all these problems, we propose a novel autocorrelation receiver equipped with a single delay-line that only has to implement a small delay. Larger delays are obtained by cycling the received signal through the delay-line and an attenuator (see Figure 1). We call this receiver the cyclic autocorrelation receiver. More specifically, we correlate $r(t)$ with the output of the cyclic delay-line:

$$r^{cd}(t) = \sum_{i=1}^{\infty} \alpha^{i-1} r(t - id), \tag{3}$$

where d is a small delay that can be implemented accurately with a small size and power consumption, and $\alpha < 1$ is the attenuation. If D_j is a multiple of d, i.e., $D_j = k_j d$, then it is clear that correlating $r(t)$ with $r^{cd}(t)$ will always include the product of $a_j h(t)$ with $\alpha^{k_j-1} h(t)$. Hence, we can compute

$$\begin{aligned} z_j^{cd} &= \int_{jT_d+D_j}^{jT_d+D_j+T_I} r(t) r^{cd}(t) dt \qquad (4) \\ &= a_j \alpha^{k_j-1} \int_0^{T_I} h^2(t) dt + \mu_j^{cd}, \end{aligned}$$

where μ_j^{cd} again includes all possible self-interference and noise terms. All the outputs z_j^{cd} are then again coherently combined to form the decision variable

$$z^{cd} = \sum_{j=0}^{N_d-1} c_j \alpha^{-k_j+1} z_j^{cd} = bN_d \int_0^{T_I} h^2(t)dt + \mu,$$

where

$$\mu = \sum_{j=0}^{N_d-1} c_j \alpha^{-k_j+1} \mu_j^{cd}.$$

An estimate of b is then again found by

$$\hat{b}^{cd} = \text{sign}(z^{cd}).$$

In general, there are more self-interference and noise terms than in the conventional autocorrelation receiver, which will degrade the performance. But as we indicated before, this structure has many benefits w.r.t. size, power consumption, and accuracy.

Still, we can advise some techniques to optimize the received signal to interference plus noise ratio (SINR). A first approach boils down to tuning α. It is clear that when α is too small, there will be no useful signal left, whereas when α is too large, the interference and noise will grow too large. Hence, there must be some optimal value of α. It is not easy to determine this analytically, due to the numerous interference and noise terms that are present in μ. We will later on show this optimal value of α by simulations.

We can further reduce the received SINR by making use of a switching device. More specifically, if we have some rough idea of where the doublet starts, we can turn the cyclic delay on when the doublet arrives and turn it off after N cycles. In that case, we correlate $r(t)$ with

$$r_N^{cd}(t) = \begin{cases} \sum_{i=1}^{n+1} \alpha^{i-1} r(t-id), \\ \text{for } nd < t \leq (n+1)d, \, n = 1, 2, \ldots, N \\ r(t-d), \\ \text{for } t \leq d \text{ and } t > (N+1)d \end{cases} \tag{5}$$

When $(N+1)d \geq D_j + T_I$, the correlator output is independent of the number of cycles N, and can be represented by

$$\begin{aligned} \tilde{z}_j^{cd} &= \int_{jT_d+D_j}^{jT_d+D_j+T_I} r(t) r_N^{cd}(t) dt \qquad (6) \\ &= a_j \alpha^{k_j-1} \int_0^{T_I} h^2(t)dt + \tilde{\mu}_j^{cd}, \end{aligned}$$

where $\tilde{\mu}_j$ is generally smaller than μ_j. The reason for this is that the number of terms in (5) is limited whereas the number of terms in (3) is unlimited. This clearly does not change the useful signal energy, but decreases the obtained interference and noise. It also means that we can take a larger value for α, without risking that the interference and noise grows unbounded. Actually, simulations show that taking $\alpha = 1$ is the best thing to do in this case.

2.4. Narrowband interference

We find it convenient and reasonable to model the narrowband interferer (NBI) $i(t)$ at UWB receiver as a single tone sinusoidal signal. Hence,

$$i(t) = \sqrt{2I} \cos(\omega_i t + \theta_i) * h_i(t), \tag{7}$$

represents the received interferer signal, with transmitted power equal to I, at the frequency $f_i = 2\pi\omega_i$, with phase θ_i. The channel $h_i(t)$ consists of the interference propagation channel $c_i(t)$ and the receive filter $q(t)$, and can generally be modeled as a frequency-flat fading channel: $h_i(t) = c_i(t) \star q(t) = \beta\delta(t-t_0)$, where β is the channel gain and t_0 the time shift. Note that the equivalent baseband model is not used and the signal is real valued. We shall merge the phase shift $\omega_i t_0$ in the phase θ_i, which can be modeled as a r.v. uniformly distributed over the interval $[0, 2\pi)$. Then, by defining $I_\beta \triangleq \beta^2 I$, we can rewrite (7) as

$$i(t) = \sqrt{2I_\beta} \cos(\omega_i t + \theta_i). \tag{8}$$

The overall received signal thus is

$$r(t) = y(t) + n(t) + i(t).$$

The term $i(t)$ will play an important role in both the conventional and cyclic autocorrelation receivers. Although NBI can be reduced by digital techniques, it is advised that the main part has already been canceled in the analog domain. Section 4 will show how to do this, but first we will look at the performance of the conventional and cyclic autocorrelation receivers without NBI.

3. SIMULATION RESULTS

In this section, we illustrate the presented ideas with a few simulation results. For simplicity, we consider a system with a single chip, i.e, $N_d = 1$. Hence, we only employ a single delay: $D_j = D$ (or $k_j = k$). A delay line that can be implemented with a reasonable accuracy is $d = 2$ ns. Hence, we will always take a D that satisfies $D = kd = 2k$ ns. As a transmit pulse, we take the first derivative of a Gaussian: $p(t) = e^{-s^2}s$, where $s = 5(t - T_p/2)/T_p$, such that the pulse has a support in $[0, T_p)$. We assume the multipath channel $c(t)$ is Rayleigh fading with $\mathcal{E}\{c(t)c(t-\tau)\} = \delta(\tau)e^{-10t/T_c}$, such that the channel has a support in $[0, T_c)$. The noise is assumed to be white Gaussian noise with power spectral density N_0. At the receiver, we filter the received signal with a filter that is matched to the transmit pulse, i.e., $q(t) = p(-t)$.

For the simulations, we consider $T_p = 1$ ns, such that the pulse $p(t)$ has a bandwidth from about 500 MHz to 2 GHz (3 dB points).

In a first experiment, we assume $T_c = 0$ ns, i.e., no multipath fading, and we compare the performance of the conventional autocorrelation receiver with the cyclic autocorrelation receiver without a switch. We take $T_I = 2$ ns, which is the length of the composite channel (this includes the propagation channel, the transmit pulse, and the receive filter). We consider $D = 6$ ns (or $k = 3$) and take different values for α. The results are presented in Figure 2. The values of α we investigate are related to the values of α^{k-1}, which represents the total attenuation of the useful signal part (see (4)). More specifically, we look at $\alpha^{k-1} = 0.4, 0.5, 0.6, 0.7, 0.8$, and 0.9. Clearly, $\alpha^{k-1} = 0.7$ is close to optimal. For other values of D (or k) a similar optimal value for α is obtained but the performance decreases with an increasing value of D (or k).

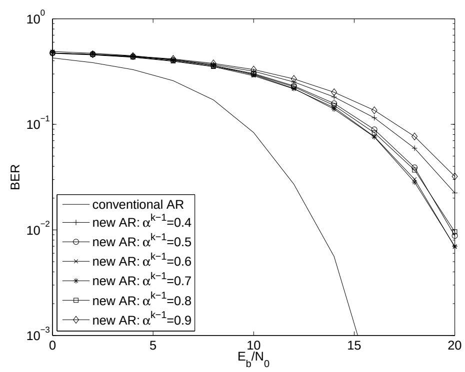

Figure 2: Performance comparison between the conventional autocorrelation receiver and the proposed one without a switch; $D = 6$ ns.

Next, let us repeat the first experiment, but let us insert a switch after the attenuation. The results are plotted in Figure 3. Clearly, the results have been improved compared to the results without a switch. Moreover, this time $\alpha = 1$ seems to be optimal, a result which holds for all values of D (or k).

Finally, let us repeat the second experiment, but let us adopt a multipath channel as defined above with $T_c = 10$ ns. The results are plotted in Figure 4. We see that the performance is still reasonable, although the multipath channel introduces some unwanted interference. Longer multipath channels could of course introduce some larger performance loss, but then more complicated equalization methods can be used.

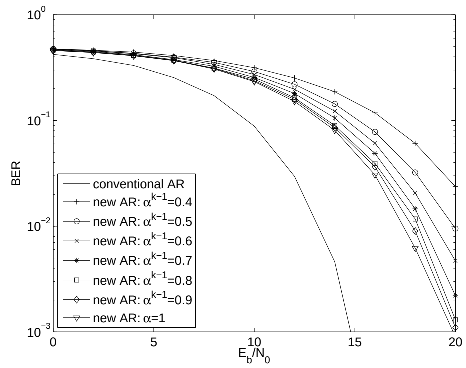

Figure 3: Performance comparison between the conventional autocorrelation receiver and the proposed one with a switch; $D = 6$ ns.

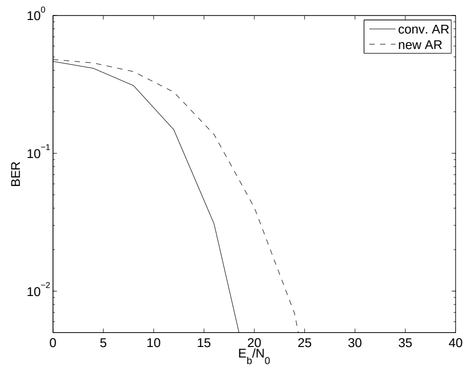

Figure 4: Performance comparison between the conventional autocorrelation receiver and the proposed one with a switch in case of a multipath channel; $D = 6$ ns.

4. SYSTEM ANALYSIS OF QUADRATURE DOWNCONVERSION AUTOCORRELATION RECEIVER

Downconversion is a technique employed in many conventional narrowband systems, as a mean of processing the information at lower frequencies and thus reducing on-chip circuit complexity and power consumption. As illustrated by Figure 5, the proposed architecture also exploits this aspect of frequency conversion. In this particular topology, the oscillator frequency is chosen such that the frequency spectrum of the incoming data is "*wrapped*" around dc, resulting in a reduced signal bandwidth without affecting neither the bit error rate (BER) nor the overall system performance. Moreover, narrowband RF frontends suffer from limited image rejection. UWB systems do not need to contend with this issue as the spectrum of a UWB signal is spread over several gigahertz and the so-called "*image*" is a fraction of the useful UWB spectrum. By frequency wrapping, interferers below 3.1 GHz are positioned at higher frequencies and thus can be easily removed by low-pass filtering of the desired signal.

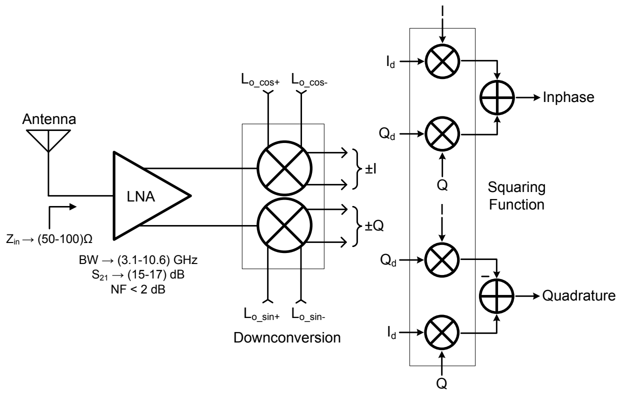

Figure 5: *Quadrature Downconversion AcR*

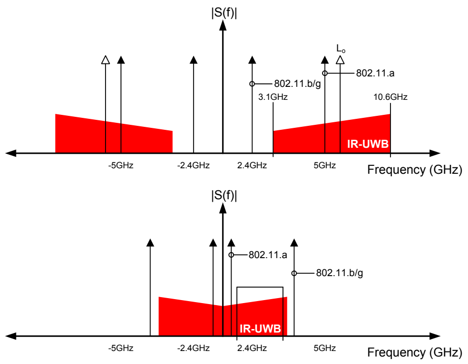

Figure 6: *Frequency Spectrum before (top) and after (bottom) downconversion*

The spectrum of the UWB signal before and after downconversion is illustrated in Figure 6.

As is seen in [4], phase and amplitude mismatch minimally affect the performance of the AcR function as compared to time delay variations, where the latter may influence the outcome of AcRs. Thus, the relative accuracy of the time delay can be considered as the figure of merit for AcRs.

In the next section, a *cyclic* delay line based on the principle of convolution is presented.

5. FILTER BASED CONTINUOUS-TIME DELAY

The QDAR resolves not only the issues such as, synchronization and capturing multipath energy, but also exploits the fact that detection with an autocorrelation function is feasible as long as the relative polarity and not the shape of consecutive pulses is preserved.

In regards to the transmitted reference systems and from an implementation point of view, one is drawn to the conclusion that the bottleneck to this concept is the physical realization of a time delay required to execute the autocorrelation function. Thus, in this section we propose a convolution based analog time delay, where the shape of the data and that of the reference pulse is traded for their relative positioning.

In this section, a procedure to derive a stable transfer function for a time delay element is discussed. One of the most important aspects of analog filter synthesis is that the approximating function must lead to a physically realizable network which is dynamically stable. There are several mathematical techniques that are frequently used to achieve the best approximation possible. A method which has proven to be successful is the Padé approximation of the Laplace transformed impulse response of the filter.

5.1. Filter Design

Once the transfer function is derived, there are many state space descriptions for a circuit that can implement it. This allows the designer to find a circuit that fits his specific requirements. In the context of low-power, low-voltage analogue integrated circuits, the most important requirements are the dynamic range, the sensitivity, and the sparsity, all of which will be treated in the subsections that follow. We will focus on a synthesis technique that is exclusively based on integrators.

5.1.1. Padé Approximant

Just like the Taylor expression, the Padé approximant is an approximation that concentrates around one point of the function that needs to be approximated. In the Padé approximation, the coefficients of the approximating rational expression are computed from the Taylor coefficients of the original function. If we were to apply the Padé approximation to $h(t)$ in the time domain, we would have to transform this function to the Laplace domain, which would possibly yield difficult expressions or even a noncausal or unstable filter [7].

The reason to apply the Padé approximation to the Laplace transform of $h(t)$ is that it immediately yields a rational expression which is suitable for implementation. Hence, a Padé approximation of $H(s)$ represents the transfer function of a possible filter. If the approximation rational function has a numerator of order *m* and a denominator of order *n*, the original function can be approximated up to order $m + n$.

Now we will derive the Padé approximation of a general function $f(t)$. Suppose that we have the Taylor series expansion of $F(s)$ around some point, e.g. $s = 0$, then

$$F(s) = c_0 + c_1 s + \ldots + c_k s^k + O(s^{k+1}) \tag{9}$$

The constants c_0 to c_k are called the Taylor coefficients of $F(s)$. Unfortunately, $F(s)$ is not a suitable expression to build a filter, since it has only zeros. Therefore, to solve this problem, we apply a Padé approximation of function $F(s)$ which is given by

$$\hat{F}(s) = \frac{P(s)}{Q(s)} = \frac{p_0 + p_1 s + \ldots + p_m s^m}{q_0 + q_1 s + \ldots + q_n s^n} \tag{10}$$

where $\hat{F}(s)$ is the truncated Taylor series given by (9) with $k = m + n$. The coefficients of $P(s)$ and $Q(s)$ can be computed as follows. The coefficients are found by setting,

$$\hat{F}(s) - \frac{P_m}{Q_n} = 0 \tag{11}$$

and equating coefficients. This procedure will yield a physically realizable transfer function.

5.1.2. Orthonormal State-Space

Among known standard state-space descriptions, such as the canonical, the diagonal and the modal, the orthonormal ladder form is notable since it is by definition semi-optimized for dynamic range due to the specific structure of the matrices. Furthermore, since it is derived from a ladder structure, it is intrinsically less sensitive and the matrices are highly sparse. A detailed explanation of the procedure to derive the orthonormal ladder form can be found in [7].

With a state-space approach, the filter can be optimized for dynamic range, sensitivity, sparsity and coefficient values. A low sensitivity suppresses the effect of component variations on the transfer function. It can be proved that a filter that is optimized for dynamic range is also optimized for sensitivity. The sparsity of the matrices directly determines the circuit complexity. State-space descriptions of filters with more zero elements require less hardware and are likely to consume lower power. Thus, it is therefore an important design aspect of state-space filters. In respect to a fully optimized and fully dense state-space description, the resulting semi-optimal orthonormal filter structure differs only by about 2dB in dynamic range. The A and b matrices of the defined transfer function are as follows,

$$A = \begin{bmatrix} 0 & \alpha_1 & & & & \mathbf{0} \\ -\alpha_1 & 0 & \alpha_2 & & & \\ & -\alpha_2 & 0 & \cdot & & \\ & & \cdot & \cdot & \cdot & \\ & & & \cdot & 0 & \alpha_{n-1} \\ \mathbf{0} & & & & -\alpha_{n-1} & -\alpha_n \end{bmatrix}$$

$$b = \begin{bmatrix} 0 \\ 0 \\ \vdots \\ 0 \\ \sqrt{\frac{\alpha_n}{\pi}} \end{bmatrix} \tag{12}$$

where all α_i's are greater than zero. If not for the single non-diagonal element, the tridiagonal A matrix would also be skew-symmetric. The b vector consists of all zeros except for the N^{th} element [8]. The *transconductance-C* or *Gm-C* technique can be used to precisely map the coefficients of the orthonormal state-space description to circuit level (see Figure 7).

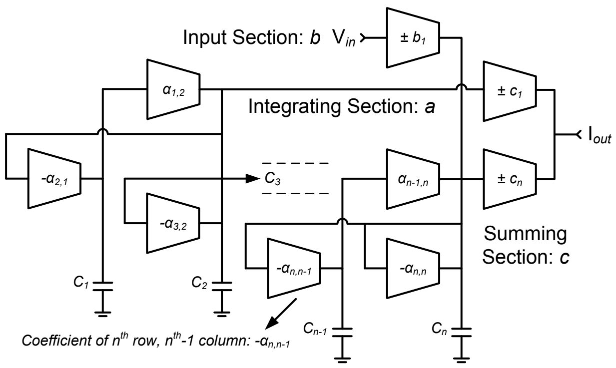

Figure 7: *Gm-C state-space filter structure for a unit time delay*

5.1.3. Transconductance Amplifier

Once the filter structure has been derived, a transconductance amplifier [9] implements every coefficient of the c vector. The orthonormal structure has both positive as well as negative coefficients, therefore a differential topology. Another advantage of using this is the cancelation of even order distortion terms that may arise from the actual nullor implementation, thus improving linearity.

5.1.4. Scaling: capacitance and coefficient values

Transconductance amplifiers will form the basic building blocks to implement the state-space description coefficients of the analog delay. The integrators are implemented as capacitors with a normalized value of 1 F. The corresponding matrix A, and vectors b and c have extremely large coefficients corresponding to large gm values, which are not physically feasible at circuit level. By scaling the capacitors and α_1, one consequently scales matrix A and vector b. Coefficients of vector c can too be down scaled by α_2, without affecting the response of the filter.

$$\hat{A} = \hat{C}A,\ \hat{b} = \alpha_1 b,\ \hat{c} = \alpha_2 c \tag{13}$$

5.1.5. Simulation Results

The group delay and the magnitude of the *cyclic* delay (i.e., a cascade of two 9^{th} order filters) are shown in Figure 8. It is clear from the plot that the group delay varies approximately $\pm 10\%$ from the mean value (i.e., 1 ns), whereas the magnitude remains relatively constant over the band of interest. By processing the received signal at lower frequencies, the QDAR's topology significantly relaxes the accuracy of the delay element. Therefore, minor fluctuations on the group delay will have little influence on the overall auto-correlation function.

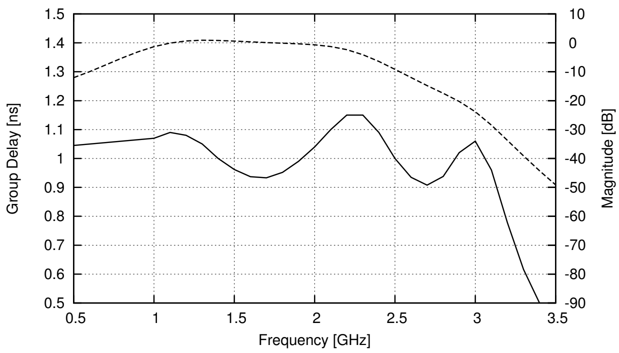

Figure 8: *Group Delay and Magnitude of the cyclic delay*

6. CONCLUSIONS

We have shown that the auto-correlation of the narrowband interferer, arising from the dirty template adopted in the signal demodulation, can be modeled. The quadrature downconversion autocorrelation receiver (QDAR) is thus be developed to counteract narrowband interferer(s) by carefully allocating them out-of-band. Moreover, this topology also circumvents processing at high frequencies, thereby relaxing the accuracy of the time delay element. Finally, we have also realized a *cyclic* delay derived from a rational transfer function of a unit delay filter.

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