ULTRASONIC ARRAY IMAGING USING CDMA TECHNIQUES

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ABSTRACT

A new method for designing ultrasonic imaging systems is presented in this paper. The method is based on the use of transducer arrays whose elements transmit wideband signals generated by pseudo-random codes, similarly to code division multiple access (CDMA) systems in communications. The use of code sequences instead of pulses, which are typically used in conventional phased arrays, combined with transmit and receive beamforming for steering different codes at each direction, permits parallel acquisition of a large number of measurements corresponding to different directions. Significantly higher image acquisition rate as well as lateral and contrast resolution are thus obtained, while axial resolution remains close to that of phased arrays operating in pulse-echo mode. Time and frequency division techniques are also studied and a unified theoretical model is derived, which is validated by experimental results.

1 INTRODUCTION

With the rapid development of ultrasonic transducer and microelectronics technologies, ultrasonic imaging has become one of the most important clinical imaging modalities. Today's state-of-the-art ultrasonic scanners involve phased array transducers of 64 to 256 elements operating in pulse-echo mode at frequencies of 2 to 20 MHz and provide gray-scale images of anatomical detail with a spatial resolution in the order of a millimeter and at a rate of more than 20 frames per second [6]. Recent advances in 2-D transducer arrays will eventually lead to the development of ultrasonic scanners capable of producing 3-D images [7].

However, 3-D image capabilities impose new requirements on the design and operation of the future ultrasonic imaging systems. Current systems typically employ 1-D arrays with up to 128 elements in order to provide increased effective aperture and thus lateral resolution. Therefore, 2-D arrays with comparable resolution would require more than 15000 elements, and even if electrical interconnection and impedance mismatching problems are solved, real-time 3-D images will only be possible with simultaneous image line acquisition [5]. Another problem with currently used techniques is that with conventional delay-and-sum beamformers grating lobes are generated which limit the dynamic range and consequently the contrast resolution [6].

In the context of this paper we present a new method for designing ultrasonic imaging systems, which employs phased-array transducers operating in continuous wave mode. Our work combines a number of techniques, namely excitation of array elements with pseudo-random code sequences, transmit and receive beamforming, as well as time and frequency division of the scanning procedure. Increased lateral and contrast resolution is achieved compared to conventional phased-array systems [3, 6]. Higher image acquisition rate is also possible due to simultaneous scanning in multiple directions, allowing real-time implementation of 3-D image generation. Earlier approaches in the use of code excitation for ultrasonic imaging include [5]. The present work, apart of providing explicit analysis of the CDMA performance using different code sequences, offers a unified model incorporating time, frequency and code division techniques combined with transmitter and receiver beamforming, allowing for optimized ultrasonic system design.

2 SYSTEM ANALYSIS

The most important part of the proposed architecture is the use of a CDMA technique for the separation of acoustic signals at different directions. A phased array transducer is employed, with different transmit and receive element spacing. Pseudo-random code sequences, or signatures, generated by finite-length shift registers [4] are used instead of pulses as input to the transmitting elements. Code sequences produce long duration, wideband acoustic signals, which can however be separated due to their autocorrelation and cross-correlation properties. Transmitted signals are then multiplexed in a way similar to multiple access environments in communication systems [1].

2.1 1-D Signal Reconstruction

In the simple case of a conventional system consisting of a single transducer element operating in pulse-echo mode and scanning one image line, reconstruction of the reflection coefficient image is based on the signal

$$\hat{R}_p(t) = \int_0^{T_m} R(\tau)p(t-\tau)d\tau \tag{1}$$

where $R(t)$ is actual reflection coefficient at depth $z = ct/2$, $p(t)$ is the pulse that is used to excite the transmitter, $T_m = 2z_{max}/c$ where z_{max} is the maximum depth of the reflecting body and c is the velocity of sound propagation in the medium [3]. In (1) attenuation is supposed to have been compensated by a gain function. On the contrary, in the case of pseudorandom code sequences, the transmitter is excited by the signal $a(t) = \sum_{j=-\infty}^{\infty} a_j p(t - jT_c)$ where $\{a_j\}$ is a pseudorandom sequence of period L and T_c is the pulse duration. In this case, reconstruction of $R(t)$ is based on the signal

$$\hat{R}_c(t) = \int_0^{T} h(u)r(t-u)du \tag{2}$$

where $T = LT_c$ is the period of $a(t)$, $h(t) = a(T-t)$, $t \in [0, T]$, is a matched filter, and $r(t) = \int_0^{T_m} R(\tau)a(t-\tau)d\tau$ is the received signal. Based on the above, one can show that (2) reduces to

$$\hat{R}_c(t) = \int_0^{T_m} R(\tau)C_a(t-\tau)d\tau \tag{3}$$

where $C_a(t)$ is the autocorrelation function of $a(t)$, and is proved to approximate a periodic delta function. Both $p(t)$ and $C_a(t)$ act as delta functions in the integrals (1) and (3). A comparison of (1) with (3) indicates that the axial resolution, which depends on the actual width of $p(t)$ or $C_a(t)$, is approximately equal in both cases. In practice, computation of the right-hand-side of (2), and consequently acquisition of $R(t)$ for all values of $z = ct/2$, is implemented by passing the received signal through a digital filter whose coefficients equal the code sequence in reverse order.

To proceed one step further, using (2) and the definitions of $a(t)$ and $r(t)$, it is possible to show that

$$\hat{R}_c(t) = \sum_{n=-1}^{M} \tilde{R}(t + (n-l)T_c)c(n-l) \tag{4}$$

where $c(n)$ is the discrete periodic autocorrelation function of sequence $\{a_j\}$, $T_m = MT_c$, $l = \lfloor t/T_c \rfloor$, $\tilde{R}(t) = \int_0^{T_m} R(\tau)C_p(t-\tau)d\tau$ and $C_p(t)$ is the autocorrelation function of $p(t)$. Equation (4) shows the reconstructed signal $\hat{R}_c(t)$ equals a sum of 'blurred' versions of $R(t)$, sampled at intervals of T_c and scaled by the autocorrelation coefficients of $\{a_j\}$. Only one term of the sum is always the desired signal, while all the others are interference terms. In the special case of maximal length sequences (or M-sequences), for which $c(n) = -1/L$ for $n \neq 0 \pmod L$, it can be shown that the interference reduces to a constant bias, which can be easily removed.

2.2 2-D Signal Reconstruction

In the more general case of a 2-dimensional reflection coefficient $R(t, \theta)$, at depth $r = ct/2$ and direction θ, several transmitted signals can be present in the body corresponding to different code sequences, in order to scan different parts of the image. Supposing that N_c sequences are used, denoted by $a_i(t)$, $i = 1, 2, \ldots, N_c$, and that no beamforming takes place (omnidirectional transmitters and receivers), reconstruction at the k-th receiver becomes

$$\hat{R}_k(t, \theta) = \sum_{i=1}^{N_c} \int_0^{T_m} R(\tau, \theta)C_{k,i}(t-\tau)d\tau \tag{5}$$

where $C_{k,i}(t)$ is the cross-correlation function between code sequences $a_k(t)$ and $a_i(t)$. Equivalently, multiple acquisitions of $R(t, \theta)$ are possible by implementing a filter bank containing the code sequences as filter coefficients. Code sequences are usually designed in such a way that $C_{k,i}(t)$ is negligible if $i \neq k$. Gold sequences are a possible candidate because of the good autocorrelation and cross- correlation properties they exhibit [2]. However, a small, suitably selected set of M-sequences is an even better solution since their autocorrelation functions are optimal (which is necessary for 1-D reconstruction) while their cross-correlation functions can be as low as those of Gold codes.

Although cross-correlation functions usually attain small values (compared to the maximum value of the autocorrelation function), the total interference represented by the sum in (5) is large enough to cause distortion of the measured reflectivity. Moreover, using several code sequences would be useless without a means to limit them to different directions. For this reason we propose transmit and receive beamforming. Several array patterns are designed for transmission, each of which corresponds to a specific code sequence and contains a main lobe steered at a certain direction. All directions θ_i, $i = 1, 2, \ldots, N_\theta$ in the region of interest $[\theta_1 \theta_2]$, are assigned a code sequence in this way, and sequences may be repeated periodically if $N_\theta = KN_c$, as described in the following section. Similar array patterns are designed for the receiver. In this case, by letting $G_i^T(\phi)$ be the i-th transmitter gain pattern and $G_\theta^R(\phi)$ a receiver gain pattern steered at direction θ, it can be shown, in a way similar to (4), that (5) becomes

$$\hat{R}_k(t, \theta) = \sum_{i=1}^{N_c} \sum_{n=-1}^{M} \tilde{R}_i(t + (n-l)T_c, \theta)c_{k,i}(n-l) \tag{6}$$

where $c_{k,i}(n)$ is the discrete periodic cross-correlation function of sequences $\{a_j^{(k)}\}$ and $\{a_j^{(i)}\}$, $\tilde{R}_i(t, \theta) = \int_{\theta_1}^{\theta_2} G_i^T(\phi)G_\theta^R(\phi)\tilde{R}(t, \phi)d\phi$ and $\tilde{R}(t, \phi)$ is defined in a

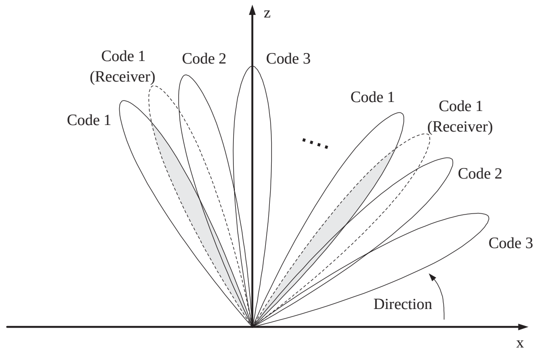

Figure 1: Illustration of different transmit and receive beamforming (side lobes are not shown in this diagram).

way similar to $\tilde{R}(t)$. In order to get the final estimation $\hat{R}(t,\theta)$ one must appropriately select those values of k in (6) which correspond to a receiver steered at direction θ, as explained in the next section. Again, one of the terms of the double summation in (6) is the desired signal, while all the others introduce interference. The n-terms correspond to interference from different depths and can reduce to a constant bias as in the 1-D case, while the i-terms correspond to interference from different directions. The latter can be removed by a correction technique which exploits the estimated signals $\hat{R}_k(t,\theta)$ to reproduce the interference terms of (6) and subtract them from the initial estimations. This technique may be applied several times and is shown in the experiments to converge to an estimation for which almost all interference is eliminated.

2.3 Transmiter and Receiver Beamforming

A typical arrangement of transmiter and receiver array gain patterns is illustrated in Figure 1. N_θ transmitter patterns with their main lobes steered at directions θ_i, $i = 1, 2, \ldots, N_\theta$ are assigned a unique code sequence, and sequences may be repeated periodically if $N_\theta = KN_c$, as in Figure 1. Similar array patterns are assigned to the receivers, but the main lobe of each receiver overlaps the two neighboring transmiter lobes. For a receiver steered at direction θ, the corresponding code sequences of the two neighboring transmiter lobes can provide us with two values of k to be used in (6) for the estimation of $\hat{R}_k(t,\theta)$. Thus, each receiver output is passed to two different filters and results in estimations for two different directions.

Gray regions of Figure 1 show such overlaps between main lobes of transmit and receive patterns for code 1. Since overlaps of main lobes corresponding to different codes (such as codes 1 and 2 in Figure 1) gives little interference (due to small cross-correlation), the above technique results in a significant increase in lateral resolution. Contrast resolution is also increased as a result of the increase in the signal-to-interference ratio. In particular, side lobes of array patterns corresponding to different directions introduce significantly lower interference to each measurement, due to the small cross-correlation between codes. Thus, multiple beam patterns are allowed to be transmitted simultaneously and all measurements are processed in parallel. Signals arriving at the receiver elements are sampled and then multiplied by different element weights (corresponding to different receive patterns and subsequently to different arrival directions) and passed through filters matched to different codes. Parallel acquisition of multiple image lines is possible in this way and this results in increased frame rate of 2-D images or real-time generation of 3-D images.

3 EXPERIMENTAL RESULTS

Figure 2 illustrates the reconstruction results of two 1-D signals (reflectivity coefficients), with and without correction. The results were obtained by using two different M-sequences of length $L = 2^9 - 1 = 511$ with low cross-correlation (actually generated by a preferred pair of primitive polynomials), and with $T_m = 63T_c$, ($M = 63$) so that $T \simeq 8T_m$ and 8 reconstructed signals fit in each sequence period. This gives us the ability to use 8 different phase shifts of each of the two primary sequences as separate code sequences for 8 directions, as explained in the next experiment. Two important conclusions can be drawn from Figure 2. (a) The reconstructed signals are obviously 'blurred' compared to the original (as shown especially for the chirp signal). This effect, which is due to the integral of Equation (5), is also present in conventional systems, and can be greatly reduced by increasing the values of M and L accordingly (in a practical system $M = 2^9 - 1 = 511$ and $L = 2^{11} - 1$ or $L = 2^{13} - 1$ would be used). (b) The 'noise' in Figure 2(b) is due to the summation of Equation (5) (actually the two received signals are added together), and can be reduced by using different phase shifts of the same sequence instead of two different sequences, resulting in much lower cross-correlation. Furthermore, this noise is almost removed by applying several iterations of the correction procedure, as shown in Figure 2(c).

Figure 3 illustrates the reconstruction results of a 2-D cyst phantom, using three primary M-sequences of length $L = 2^{11} - 1 = 2047$, $N_\theta = 24$ trasmitter array gain patterns with 24 separate receiver patterns, and $T_m = 255T_c$ ($M = 2^8 - 1$). The 24 code sequences used for the array patterns were obtained by using 8 phase shifts of each primary sequence. This is possible since $T \simeq 8T_m$ and results in acquiring 8 image lines at each sequence period. Thus, reconstruction results are produced at different time 'slots' for each direction, resulting in a kind of time-division multiple access (TDMA). Note that severe artifacts are present in Figure 3(c) at the two boundaries of the three primary sequences, due to their higher cross-correlations, but can be removed with the correction procedure (Figure 3(d)). Finally, a conventional system employing only receiver beam patterns obviously gives a more 'blurred' reflectivity estimation (Figure 3(b)).

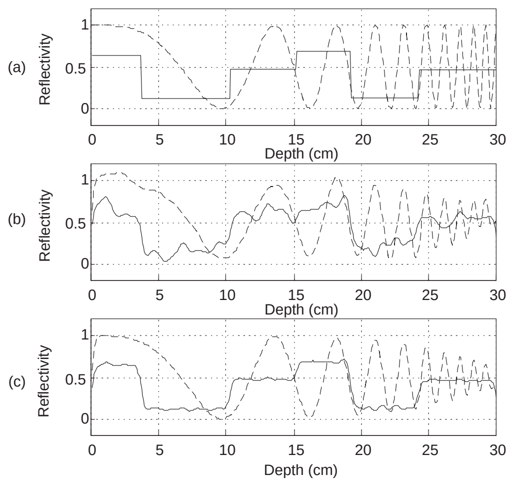

Figure 2: (a) Original reflectivity coefficients, (b) reconstruction without correction, and (c) with correction.

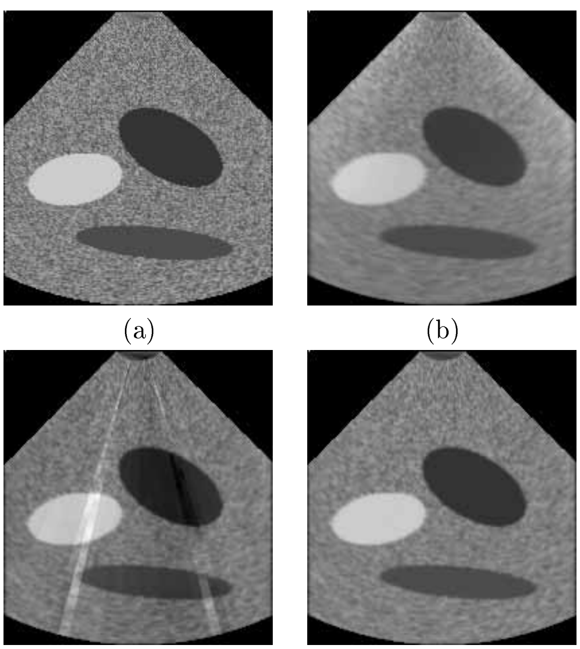

Figure 3: (a) Original 2-D reflectivity coefficient (cyst phantom), (b) conventional reconstruction, (c) reconstruction using the proposed method without correction, and (d) with correction

4 DISCUSSION-CONCLUSIONS

It is evident from the experimental results that time division of the acquisition procedure can be achieved, by dividing image acquisition in $N_t = M/L = T/T_m$ time slots so that N_θ/N_t measurements are performed in each slot. Higher contrast resolution is achieved in this way at the additional cost of slower frame rate, since cross-correlation values are then scaled by a factor of $1/N_t$. Similarly, transmitted and received signals can be divided in N_f frequency bands if the total transducer bandwidth allows that. Again total interference is reduced at the cost of shorter code sequences and thus higher cross-correlation functions.

Useful conclusions can be drawn from the theoretical model presented in this paper, concerning the selection of parameters such as N_c, N_θ, N_t, N_f, K, L, number and locations of transmitter elements and beam patterns. These selections depend upon the designed system characteristics, such as desired spatial and contrast resolution, acquisition speed, total available transducer bandwidth, operating frequency etc. Using the proposed model the designer of the overall system may systematically tune its parameters taking into account the tradeoffs between the performance of the system in terms of axial and lateral resolution, SNR levels and scanning speed.

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