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# A RANDOM SIGNAL ULTRASONIC NDT SYSTEM FOR HIGHLY ATTENUATING MEDIA

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## Abstract

This paper presents a novel radiofrequency ultrasound pulse-echo or pulse-transmission system for materials evaluation or flaw detection. It is specifically designed to provide wide bandwidth operation in very lossy materials such as rubbers and many engineering polymers. Instead of the conventional short pulse or gated sinewave burst the system transmits pseudo random binary sequences (PRBS) which have broad spectra, high average and low peak energy. Acoustic pathway characterisation or flaw detection is carried out by a cross-correlation operation between transmitted and received sequences, using a fast analogue multiplier. The result is improvements in signal to noise ratio at high frequencies that could only be achieved otherwise by fast averaging hardware, at much greater cost and slower repetition rate. The PRBS clock rate achieved in our prototype is 20 MHz, giving an effective test bandwidth of 10 MHz.


## Introduction

Many polymeric engineering materials absorb ultrasound at a rate that increases rapidly with frequency. Signal components of short wavelengths are therefore masked by electronic noise in test systems and this limits the minimum size of defect that can be detected nondestructively. The performance of conventional ultrasonic flaw detectors under these poor signal to noise conditions can be improved by coherent averaging of successive A-scans. This is time consuming, costly in terms of computational effort, and the result may still be limited by poor matches between the wave shape of the transmitter excitation and the transducer impulse response. An alternative method of signal to noise ratio improvement is described here; it is based on the transmission through the test medium of pseudo-random signals of low amplitude, long duration and wide bandwidth, combined with signal pathway identification using a correlation method. This paper gives an outline of the method and presents preliminary results obtained using a practical system.

Signal path identification using random signals

A linear time-invariant (LTI) signal path can be characterised by its impulse response h(t) in the time domain or equivalently by its frequency response in the frequency domain. For input x(t), the output y(t) is obtained by convolution, thus:

$$y(t) = \int x(t - \tau)h(\tau)d\tau \qquad (1)$$

The equivalent frequency domain operation is:

$$Y(\omega) = X(\omega)H(\omega) \qquad (2)$$

where upper case symbols X, Y, H are the Fourier transforms of the lower case time domain symbols, x, y, h. The cross-correlation function expressing output to input similarity is:

$$r_{xy}(\tau) = \frac{1}{T}\int_{-T}^{T} x(t + \tau)y(t)dt \qquad (3a)$$

In the frequency domain this operation is equivalent to the following multiplication of transforms:

$$P_{xy}(\omega) = X^{*}(\omega)Y(\omega) \qquad (3b)$$

where * indicates complex conjugate and $P_{xy}(\omega)$ and $r_{xy}(\tau)$ form a Fourier transform pair. Combining (2) and (3) we get:

$$P_{xy}(\omega) = X(\omega)X^{*}(\omega)H(\omega) \qquad (4)$$

Now the product $X(\omega)X*(\omega)$ is the Fourier transform of the autocorrelation function of the input signal, given by:

$$r_{xx}(\tau) = \frac{1}{T}\int_{-T}^{T} x(t + \tau)x(t)dt \qquad (5)$$

and

$$\int_{-\infty}^{\infty} r_{xx}(\tau)e^{j\omega\tau}\, d\tau = X^{*}(\omega)\, X(\omega) \qquad (6)$$

Now transforming (4) back into the time domain we get

$$r_{xy}(\tau) = \int_{-\infty}^{\infty} r_{xx}(\tau - T)\, h(T)dT \tag{7}$$

The meaning of this integral is that the cross-correlation function between output and input of a LTI pathway is equal to the convolution of the input autocorrelation function with the pathway impulse response. If the input signal is random and spectrally white its auto-correlation function will be a unit impulse. The result of implementing the cross-correlation of (3) will be the impulse response of the propagation path. The advantage of using random signals instead of short pulses for test radiation in NDT of lossy pathways is that more energy at high frequencies can be injected from the source than would be possible by increasing the short pulse voltage drive to a transducer up to the maximum electric field limit of the device. With extra signal processing it is also possible to prefilter transmitted white noise to emphasise high frequencies and thereby reduce the dynamic range of signals that arrive at the receiver after propagation through pathways exhibiting high frequency loss. The effectiveness of the random noise technique in improving signal to noise ratio can be demonstrated by considering the input-output relationships. For a system which generates additive noise, fig 1, if the signal at the system output is s(t) and the additive noise is n(t) the input-output cross-correlation is

$$r_{xy}(\tau) = \frac{1}{T} \int_{-T}^{T} x(t+\tau) \left[s(t) + n(t)\right] dt \tag{8}$$

or

$$r_{xy}(\tau) = \frac{1}{T} \int_{-T}^{T} x(t+\tau)s(t)dt + \frac{1}{T} \int_{-T}^{T} x(t+\tau)n(t)dt \tag{9}$$

Now in well designed electronic hardware the additive noise n(t) is uncorrelated to both output and input; the second term in (9) therefore reduces to zero provided that the interval of integration, T, is large enough. The first integral in (9) is the cross-correlation between system output and input uncorrupted by noise. As above, if x(t) is truly random and spectrally white, then $r_{xy}(\tau)$ reduces to the required impulse response h(t). The correlation method will therefore remove the effect of system noise from the estimate of h(t).

Practical generation of white noise

Truely white noise cannot be generated in a practical system because infinite bandwidths are not available. Similarly truely random processes are only so when assessed over infinite time intervals. "White" noise in a practical system is therefore bounded in both time and frequency domains and is not ideally random. The most convenient way to generate noise-like signals is to use shift register technology, or its software counterpart, to produce pseudo-random binary sequences (PRBS) [1,2,3]. These are binary signals with random switching times between binary states. They have structure in that successive states are generated at a frequency governed by a system clock. The ideal PRBS has an auto-correlation function (ACF), when expressed in continuous lag-time, that is triangular in form and of two clock pulses duration. Such signals could, in principle, be used to identify h(t) over a bandwidth between dc to just below half the clock frequency. However, simple PRBSs generated in practical systems display a periodic "tail" at non-zero lags in their ACFs and this can lead to serious errors in estimates of h(t). The problem can be overcome by using complementary pairs of PRBSs discussed by Golay[4].

Golay Sequences

Golay defined a set of complementary series as a pair of equally long finite sequences of two kinds of elements which have the property that the number of pairs of like elements with any given separation in one series is equal to the number of pairs of unlike elements with the same separation in the other series. One important outcome of this property is seen in the sequence auto-correlations. Both have a triangular spike two clock pulses wide at zero lag-time and both have low level oscillations at longer lags. However the oscillations in the ACF of one sequence are in antiphase with the oscillations in the ACF of the complementary sequence and the sum of the two ACFs results in the ideal triangular function at zero lag with no significant finite energy at non-zero lags. In order to identify the impulse response h(t) in the signal scheme of fig 1 it is necessary to transmit each Golay sequence separately through the system, compute the output-input cross-correlation functions for each sequence, and then add these together. If $G_A(t)$ and $G_B(t)$ are the complementary sequences then the summed cross correlograms become:

$$R_H(\tau) = \int_t G_A(t+\tau) \int_T G_A(t-T)h(T)dTdt + \int_t G_B(t+\tau) \int_T G_B(t-T)h(T)dTdt + \int_t \left[G_A(t+\tau) + G_B(t+\tau)\right] n(t)dt \quad (10)$$

It can be shown that the correlation and convolution operations implicit in 10 can be rearranged to give:

$$R_H(\tau) = \int_T \left[R_A(\tau - T) + R_B(\tau - T)\right] h(T)dT + \int_T \left[G_A(t+\tau) + G_B(t+\tau)\right] n(t)dt \quad (11)$$

Here $R_A$ and $R_B$ are the ACFs of the complementary sequences $G_A$ and $G_B$, defined above. Equation 11 is to be interpreted as the sum of the complementary sequence autocorrelations in convolution with the system impulse response plus the sum of the original complementary sequences in cross-correlation with the uncorrelated system noise n(t). The second term approximates to zero, leaving the first term which provides the required impulse response, identified over a bandwidth up to just less than half the sequence clock frequency.

Prototype system

Fig 2 shows a block diagram of our prototype system. It is designed to operate at sequence clock rates up to 20 MHz, although we have found twice that speed possible. The system communicates with a host computer by 16 bit parallel cable and a single analogue input port. The system consists of two identical banks of memory each being divided into two sections, one for sequence A and one for its complementary sequence B. At the start of a run both banks are loaded from the host computer, as is a set of control codes that set the clock rate and sequence length, upto a maximum of 8K successive binary states. The A-sequence is first transmitted through the test sample and an identical "copy" from the second memory bank is used to estimate the cross correlation between sequence A and sequence A after transmission through the test sample. The process is repeated for sequence B. The two output cross-correlation functions are then added together in the host computer to yield the estimate of the signal pathway h(t) that is required.

In this system the binary sequences are output from an 8-bit digital memory via a fast digital to analogue converter (DAC), rather than a simple single bit binary output stage that would have satisfied the basic requirement. The reason that we have chosen the more complex structure is to enable us to precolour the transmitted sequences in order to whiten the spectrum of signals that are received from the test propagation path.

Preliminary results

Two 1.25 MHz centre frequency highly damped commercial flaw detector transducers were arranged coaxially on either side of a mild steel block of thickness 60 mm. A 60 dB electrical attenuator was placed between the receiving transducer and the receiver amplifier to bring the total path loss to 95 dB. Another attenuator was placed between the flaw detector pulse output and the transmitting transducer in order to bring the flaw detector's transmission signal down to 20V peak, to match the amplitude of the Golay sequence. The signal received from the conventional flaw detector is shown in fig. 3, whilst that from the same pathway, but obtained using the Golay sequence is shown in fig. 4. It is clear that the new method results in a dramatic improvement in signal to noise ratio and enables successful identification of high loss signal paths that would otherwise be masked by system noise.

Concluding remarks

This paper has demonstrated a pseudo-random noise technique for the evaluation of acoustic signal pathways in lossy media. Use is made of Golay complementary series to overcome limitations of the simple PRBS. Whilst further evaluation is still necessary the signal to noise performance of the system can be seen to exceed that of the coherent average method provided that test sequences of adequate length are used.

References

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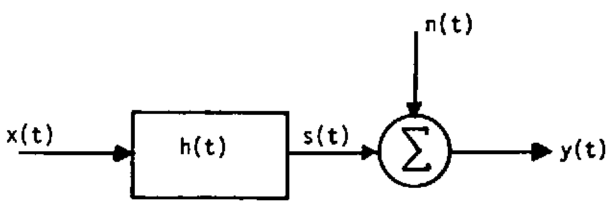


Fig 1 Signal paths in an LTI system with additive noise.

Block diagram of prototype

Golay sequencer system.

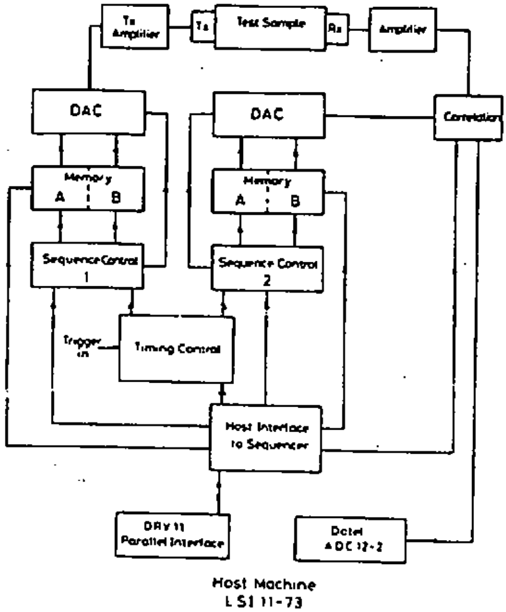


Fig 2

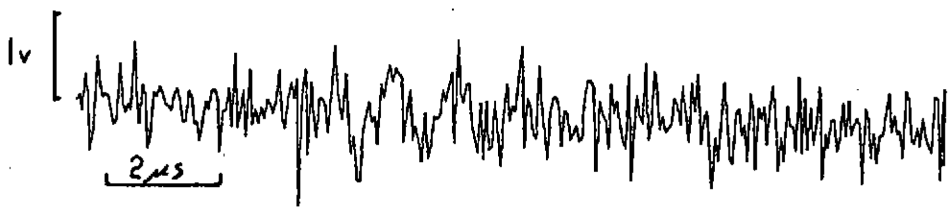


Fig 3

Signal obtained from 95 dB loss transmission path using a conventional flaw detector.


Fig 4 Signal obtained as estimate of h(t) for the same transmission path, using the Gólay sequencer.