



M U H A M M A D T A Y Y I B

**S P R E A D S P E C T R U M
S I G N A L S A P P L I C A T I O N
I N R E S O N A N T
U L T R A S O N I C
S P E C T R O S C O P Y
M E A S U R E M E N T S**

D O C T O R A L D I S S E R T A T I O N

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MUHAMMAD TAYYIB

SPREAD SPECTRUM SIGNALS APPLICATION
IN RESONANT ULTRASONIC
SPECTROSCOPY MEASUREMENTS

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CONTENT

LIST OF FIGURES.....	7
LIST OF TABLES.....	14
LIST OF ABBREVIATIONS.....	15
INTRODUCTION	16
1. PROBLEM ANALYSIS.....	24
1.1. Time of flight estimation	24
1.2. Thickness resonant ultrasound spectroscopy	25
1.3. Acoustic wave propagation	28
1.4. Ultrasonic transducers	31
1.5. Excitation signals for ultrasonic measurements	33
1.5.1. Bipolar and unipolar excitation signals for transducers	36
1.6. Conclusions	38
2. DESIGN AND OPTIMIZATION OF APWP SIGNALS FOR TRANSDUCTION SPECTRAL LOSSES COMPENSATION	39
2.1. TRUS for thin sample in air	40
2.2. The need for high SNR and wide bandwidth in TRUS	42
2.3. Unipolar excitation for flat SNR	45
2.4. SNR flatness evaluation on TRUS	51
2.4.1. Experimental setup: excitation and reception electronics.....	52
2.5. Sample parameters estimation performance	55
2.6. Conclusions	58
3. STUDY OF SPECTRAL CONTENT EFFECTS IN NON-CONTACT RESONANCE ULTRASOUND SPECTROSCOPY	59
3.1. Air-coupled resonance ultrasound spectroscopy	59
3.2. APWP signals.....	61
3.2.1. Frequency dependent noise	62
3.2.2 SNR equalization.....	62
3.2.3. APWP signals for optimal spectrum shaping	63
3.3. Experimental setup	67
3.4. Results	68
3.4.1. Spectrum location and bandwidth influence.....	68
3.5. Conclusions	71
4. SOLUTION FOR LONG SIGNALS OVERLAPPING.....	72
4.1. Overlapping in long duration excitation signals	72
4.2. Deep neural network approach for overlapping solutions	74
4.2.1. Res-UNet1D based DNN mode.....	76
4.2.2 Res-UNet1D model results for long duration signals.....	78
4.3. Deconvolution for Long Spread Spectrum Signals	79
4.3.1. Problem analysis.....	82
4.3.1.1. Pulse compression using standard approach.....	83
4.3.1.2. Pulse compression using inverse filter	86
4.3.1.3. Pulse decompression	87
4.3.2. Performance evaluation	87

4.3.2.1. Simulation results	88
4.3.2.2. Experimental results	93
4.4. Conclusions	95
5. SPERFORMANCE EVALUATION IN THICKNESS RESONANCE	
ULTRASOUND SPECTROSCOPY	97
5.1. Transducers used	97
5.2. Experiment setup	98
5.3. Results of air-coupled NC-RUS measurements	99
5.4. Results of immersion TRUS measurement.....	106
5.5. Conclusions	109
CONCLUSIONS.....	111
SANTRAUKA.....	112
LIST OF REFERENCES.....	137
CURRICULUM VITAE AND DESCRIPTION OF CREATIVE ACTIVITIES (CV).....	149
LIST OF SCIENTIFIC PAPERS AND SCIENTIFIC CONFERENCES.....	151

LIST OF FIGURES

Fig. 1 (left) Partially overlapping reflections and (right) isolated plot of the 1st and 2nd reflections, shown separately for clarity.....	24
Fig. 2. (left) Fully overlapping reflections and (right) isolated plot of the 1st and 2nd reflections.....	25
Fig. 3. (left) Time domain overlapping of reflections for noise free signal, (right) Wiener deconvolution (regularization parameter = 10^{-12}) when reflections are overlapping for noise free signal.....	25
Fig. 4. Effect on transmission response of (top left) sample attenuation, (top right) velocity, (bottom left) density and (bottom right) thickness.....	26
Fig. 5. Transmission response of (left) organic and (right) non-organic samples.....	27
Fig. 6. Frequency-dependent sound attenuation in air at 20°C, 1 atm, with 50% relative humidity and 10 mm distance path, (left) attenuation over 2 MHz frequency range, (right) attenuation over 200 kHz frequency range.....	28
Fig. 7. (left) Atmospheric pressure vs. (right) temperature and relative humidity influence on speed of sound.....	29
Fig. 8. Air-solid interface using air-coupled ultrasonic transducer.....	30
Fig. 9. Schematic of transducer.....	31
Fig. 10. Transmission response of $\lambda/4$ matched PZT (left) 650 kHz center frequency transducer, (middle) 1 MHz center frequency transducer and (right) capacitive transducer pair [140]	32
Fig. 11. (left) Sinusoidal chirp LFM in time and (right) frequency domain.....	35
Fig. 12. Autocorrelation of sinusoidal chirp and tapered chirp for sidelobe compression.....	35
Fig. 13. (left) Transmitted 100 μ s LFM signal and (right) received signal after passing through system model for unipolar and bipolar excitation cases.....	36
Fig. 14. SNR of received signal using 100 μ s LFM signal after passing through system model for unipolar and bipolar excitation cases.....	37
Fig. 15. TRUS in measurement modes: (left) calibration and (right) sample.....	41
Fig. 16. Transmission response of experimentally measured <i>Vitis vinifera</i> leaf, (left) magnitude, and (right) phase response	41
Fig. 17. The magnitude spectrum of the received signal when pulse (red) is used as excitation signal and LFM (black) signal for air-coupled transducer.....	43
Fig. 18. The magnitude spectrum of the received signal when air-coupled transducer is excited by LFM (red) and losses-compensating APWP signal (black) signal.....	43
Fig. 19. Preamplifier input NVSD when connected to receiving transducer.....	44
Fig. 20. Received signal SNR when transducer is excited by LFM (red) and losses compensating APWP (black) signal.....	44
Fig. 21. A unipolar excitation signal for APWP.....	45

Fig. 22. SNR compensation for spectral losses signal path.....	45
Fig. 23. Compensation function.....	47
Fig. 24. (left) Time-frequency characteristic function of compensated NLFM signal, (right) interpolated instantaneous frequency function.....	47
Fig. 25. (left) Comparison of transmitted and (right) received signals spectra for the desired (blue), sinusoid NLFM (black) and rectangular NLFM (red) signals after spectral losses compensation. Note that the desired spectral shape was not obtained.....	47
Fig. 26. (left) Time domain unipolar NLFM signal of 100 μ s, (center) time domain unipolar APWP signal of 100 μ s, (right) time domain segment of APWP pulse train for unipolar case: black not compensated, blue compensated.....	50
Fig. 27. (left) SNR spectral flatness improvement ($\Delta\xi$) for each iteration and (right) total improvement (ξ) in each iteration.....	50
Fig. 28. The received SNR comparison. Note that SNR is flat only for APWP signal. Bandwidth was improved for NLFM and APWP signals.....	51
Fig. 29. Transmission magnitude for <i>Vitis Vinifera</i> leaf in case of different excitation signals: (top left) pulse, (top right) LFM, (bottom left) NLFM and (bottom right) APWP	52
Fig. 30. Experimental setup used for the evaluation of polycarbonate sample by using air-coupled ultrasonic data acquisition system.....	53
Fig. 31. Low-noise inverting preamplifier using LMH6624MA with capacitive stabilization [35]	53
Fig. 32. System-level block diagram of a portable air-coupled ultrasonic measurement system [29]	54
Fig. 33. Bipolar pulser using high and low side driver [54].....	54
Fig. 34. Transmission magnitude for polycarbonate plate in case of different excitation signals: (top left) pulse, (top right) LFM, (bottom left) NLFM and (bottom right) APWP.....	55
Fig. 35. Relative error attenuation (left) <i>Vitis vinifera</i> and (right) polycarbonate....	56
Fig. 36. Relative error velocity (left) <i>Vitis vinifera</i> and (right) polycarbonate.....	56
Fig. 37. Relative error density (left) <i>Vitis vinifera</i> and (right) polycarbonate....	57
Fig. 38. Relative error thickness (left) <i>Vitis vinifera</i> and (right) polycarbonate.....	57
Fig. 39. Air-coupled measurement setup.....	60
Fig. 40. Time domain TRUS fitting result for sample signal using pulse excitation: model: (left) signal comparison and (right) remainder after model signal subtraction from experimental one	61
Fig. 41. Received signal after pulse excitation: (left) time domain, (right) normalized in spectral domain (circles: -30 dB bandwidth)	61
Fig. 42. Compensation function for SNR.....	62
Fig. 43. Pseudocode for SNR based spectral compensation.....	64
Fig. 44. (right) Energy of excitation signal compared to (left) average SNR of received sample signal for broadest bandwidth coverage, 650 kHz (100% of center frequency)	65

Fig. 45. (right) Energy of excitation signal compared to (left) average SNR of received sample signal for broadest bandwidth coverage, 400 kHz (60% of center frequency)	66
Fig. 46. (right) Energy of excitation signal compared to (left) average SNR of received sample signal for broadest bandwidth coverage, 100 kHz (15% of center frequency)	66
Fig. 47. Experimental measurement setup: (top) equipment view with insertion mechanism removed; (bottom) insertion mechanism view	67
Fig. 48. Transmission response of polycarbonate sample: (left) magnitude, (right) phase	68
Fig. 49. Attenuation a bias error for the polycarbonate sample when the excitation coverage is concentrated (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies	69
Fig. 50. Ultrasound velocity c bias errors for the polycarbonate sample when the excitation coverage is (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies	69
Fig. 51. Density ρ bias errors for the polycarbonate sample when the excitation coverage is (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies	70
Fig. 52. Thickness h bias errors for the polycarbonate sample when the excitation coverage is (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies	70
Fig. 53. Spectral response of compensation SNR using pulse, NLFM, LFM, ζ_{APWP} , and ζ_{LFMdB} . (pulse signal is not compensated)	71
Fig. 54. Calibration signal (AC response of transducer) of (left) 30 μ s, (middle) 100 μ s and (right) 250 μ s long LFM excitation signal	72
Fig. 55. Sample signal (sample inserted between transducers in through transmission of air-coupled ultrasound) of LFM with (left) 30 μ s, (middle) 100 μ s, (right) 250 μ s excitation signal duration	73
Fig. 56. Transmission response of (left) <i>vitis vinifera</i> (grape plant leaf) when reflections are separate, (right) when reflections are overlapping	73
Fig. 57. Spectral response comparison between LFM and pulse excitation signals	74
Fig. 58. (top) Temporal response, (bottom) spectral response of 50 μ s chirp compared with 200 μ s chirp	75
Fig. 59. Res-UNet1D based methodology flow diagram	76
Fig. 60. Architecture of the 1D convolutional autoencoder model	77
Fig. 61. (left) Temporal, (center) spectral and (right) residual response of 50 μ s that was part of the training	77
Fig. 62. (left) Temporal, (center) spectral and (right) residual response of 250 μ s that was part of the training	78

Fig. 63. (left) Temporal, (center) spectral and (right) residual response of 350 μ s that was part of the training.....	78
Fig. 64. The trade-off between resolution, energy, and bandwidth in signal design.....	79
Fig. 65. Autocorrelation of (left) pulse, (center) LFM and (right) NLFM signals.....	80
Fig. 66. Calibration signals AC response of three different center frequency transducers i.e. 300 kHz, 650 kHz and 1 MHz.....	81
Fig. 67. Calibration signal (left) time domain and (right) frequency domain AC response in case of wideband 100 ns pulse excitation. Bandwidth at -6 dB and -20 dB.....	82
Fig. 68. (left) Calibration signal using LFM and (right) corresponding pulse compression using matched filter, red comparison to pulse (scaled).....	83
Fig. 69. (left) Comparison of pulse excitation and (right).MF compressed LFM. Compressed signals were normalized.....	83
Fig. 70. Mismatched, Blackman-Harris filter-based, pulse compression when reference signal is: (left) excitation code and (right) calibration signal.....	84
Fig. 71. Sample measurements using (left) 100 ns pulse excitation and (right) 50 μ s. LFM excitation.....	84
Fig. 72. (left) Spectra of calibration (red), sample (black) and noise (blue) signals vs. (right) measured (red) and fitted (black) sample transmission response.....	85
Fig. 73. Comparison of sample signal pulse compression result using (left) MF and (right) MMF. Signals normalized to pulse.....	85
Fig. 74. Sample signal contains overlapping reflections: (left) explanation of propagation; (right) pulse compression result using MF.....	85
Fig. 75. (left) calibration signal using 350. 350 μ s LFM signal with inverse filter using pulse compression, (right) corresponding sample transmission.....	86
Fig. 76. (left) calibration signal using 350. 350 μ s LFM signal with inverse filter using pulse compression, (right) corresponding sample transmission.....	87
Fig. 77. Hilbert envelope of (left) calibration and (right) sample signals in case of 50 μ s duration LFM excitation.....	88
Fig. 78. Hilbert envelope of (left) calibration and (right) sample signals in case of 100 μ s duration LFM excitation.....	89
Fig. 79. Hilbert envelope of (left) calibration and (right) sample signals in case of 350 μ s duration LFM excitation.....	89
Fig. 80. Transmission response of (left) sample 10 μ s 10 V LFM, (right) 350 μ s 50 V LFM.....	89
Fig. 81. Parameter excitation of NC-RUS for 10 μ s 10 V LFM.....	90
Fig. 82. LFM10 μ s attenuation a bias.....	90
Fig. 83. Bias errors for 100 μ s 10 V LFM.....	91
Fig. 84. Bias errors for 350 μ s 10 V LFM.....	91

Fig. 85. (left) Spectral losses compensated spectra of calibration and sample with 50 V APWP of 350 μ s duration, (right) transmission match with 50 V APWP of 350 μ s duration.....	92
Fig. 86. (left) Spectral losses compensated for AC response of sample transmission with 50 V APWP of 350 μ s duration, (right) transmission match with 50 V APWP of 350 μ s duration.....	92
Fig. 87. Bias errors for 350 μ s 50 V APWP.....	93
Fig. 88. Bias errors for transmission match 350 μ s 50 V APWP.....	93
Fig. 89. Bias errors for 10 μ s 20 V LFM.....	94
Fig. 90. Bias errors for 100 μ s 20 V LFM.....	95
Fig. 91. Bias errors for 350 μ s 20 V LFM.....	95
Fig. 92. Transducers used in investigation: (left) air-coupled and (right) immersion.....	97
Fig. 93. AC transmission response: (left) air-coupled and (right) immersion.....	97
Fig. 94. Spectroscopy measurement setup: (left) air-coupled and (right) immersion.....	98
Fig. 95. Excitation signals used in air-coupled NC-RUS measurements spectra for (left) wideband and (right) narrowband set.....	100
Fig. 96. Received (left) calibration and (right) sample signal spectra.....	100
Fig. 97. Sample transmission response (left) magnitude and (right) phase for wideband LFM (label WB) at 50 V excitation.....	101
Fig. 98. Sample transmission response (left) magnitude and (right) phase for LFM bandwidth matched to -20 dB passband (label ch20dB) at 50 V excitation.....	101
Fig. 99. Sample transmission (left) magnitude and (right) phase for APWP with losses compensation at -20 dB passband (label AW20dB) at 50 V excitation.....	101
Fig. 100. Sample transmission response (left) magnitude and (right) phase for LFM bandwidth matched to -6 dB passband (label ch6dB) at 50 V excitation.....	102
Fig. 101. Sample response (left) magnitude and (right) phase for APWP with losses compensation within -20 dB passband (label AW6dB) at 50 V excitation.....	102
Fig. 102. Sample transmission response (left) magnitude and (right) phase for wideband LFM (label WB) at 100 V excitation.....	102
Fig. 103. Sample transmission response (left) magnitude and (right) phase for LFM bandwidth matched to -20 dB passband (label ch20dB) at 100 V excitation.....	103
Fig. 104. Sample transmission (left) magnitude and (right) phase for APWP with losses compensation at -20 dB passband (label AW20dB) at 100 V excitation.....	103
Fig. 105. Sample transmission response (left) magnitude and (right) phase for 500 ns pulse (label Pu500ns) at 100 V excitation.....	103
Fig. 106. (left) Sample magnitude transmission response for wideband LFM (label WB) and (right) for APWP with losses compensation at -20 dB passband (label AW20dB) at 20 V excitation.....	104
Fig. 107. NC-RUS errors for (top left) sample thickness h , (top right) velocity c , (bottom left) density and (bottom right) attenuation at 100 V excitation.....	104
Fig. 108. NC-RUS errors for sample (top left) sample thickness h , (top right) velocity c , (bottom left) density r and (bottom right) attenuation a at 20 V excitation.....	105
Fig. 109. Bias NC-RUS errors for sample thickness h estimation at (left) 20 V excitation, (center) 50 V excitation and (right) 100 V excitation.....	105

Fig. 110. Random NC-RUS errors for sample thickness h estimation at (left) 20 V excitation, (center) 50 V excitation and (right) 100 V excitation.....	106
Fig. 111. Excitation signals used in immersion TRUS measurements spectra for (left) wideband and (right) narrowband set.....	106
Fig. 112. Received (left) calibration and (right) sample signal spectra.....	106
Fig. 113. Sample transmission magnitude: (left) wideband LFM, (center) LFM -20 dB passband and (right) APWP with -20 dB passband	107
Fig. 114. Sample transmission magnitude: (left) LFM -6 dB passband, (center) APWP -20 dB passband and (right) 100 ns pulse.....	107
Fig. 115. TRUS errors for sample (top left) thickness h , (top right) velocity c , (bottom left) density r and (bottom right) attenuation α in immersion measurements.....	108
Fig. 116. Bias TRUS errors for sample (top left) thickness h , (top right) velocity c , (bottom left) density r and (bottom right) attenuation α	108
Fig. 117. Random TRUS errors for sample (top left) thickness h , (top right) velocity c , (bottom left) density r and (bottom right) attenuation α	109

PAVEIKSLŲ SĄRAŠAS

118 pav. Priimto signalo amplitudės spektro (kairėje) ir SNR (dešinėje)	122
119 pav. Vienpolinis sužadinimo signalas APWP.....	122
120 pav. perdavimo dydis <i>Vitis Vinifera</i> lapui esant skirtingiems sužadinimo signalams: impulsui (viršuje kairėje), LFM (viršuje dešinėje), NLFM (apačioje kairėje) ir APWP (apačioje dešinėje)	123
121 pav. Santykinė paklaida a : <i>Vitis vinifera</i> (kairėje) ir polikarbonato (dešinėje)	124
122 pav. Santykinė paklaida c : <i>Vitis vinifera</i> (kairėje) ir polikarbonato (dešinėje)	124
123 pav. Santykinė paklaida r <i>Vitis vinifera</i> (kairėje) ir polikarbonato (dešinėje) ...	124
124 pav. Santykinė paklaida h <i>Vitis vinifera</i> (kairėje) ir polikarbonato (dešinėje) ..	125
125 pav. Polikarbonato mėginio silpnėjimo nuokrypio klaidos, kai sužadinimo aprėptis yra sutelkta aplink rezonanso piką (kairėje), plečiasi nuo rezonanso piko link aukštesnių dažnių (viduryje) ir plečiasi nuo rezonanso piko link žemesnių dažnių (dešinėje)	127
126 pav. Ultragarso greičio nuokrypio paklaidos polikarbonato mėginyje, kai sužadinimo aprėptis yra (kairėje) sutelkta aplink rezonanso piką, (viduryje) plečiasi nuo rezonanso piko link aukštesnių dažnių ir (dešinėje) plečiasi nuo rezonanso piko link žemesnių dažnių.....	127
127 pav. Polikarbonato mėginio tankio nuokrypio paklaidos, kai sužadinimo aprėptis yra (kairėje) sutelkta aplink rezonanso piką, (viduryje) plečiasi nuo rezonanso piko link aukštesnių dažnių ir (dešinėje) plečiasi nuo rezonanso piko link žemesnių dažnių.....	128
128 pav. Polikarbonato mėginio storio nuokrypio paklaidos, kai sužadinimo aprėptis yra (kairėje) sutelkta aplink rezonanso piką, (viduryje) plečiasi nuo rezonanso piko link aukštesnių dažnių ir (dešinėje) plečiasi nuo rezonanso piko link žemesnių dažnių.....	128

129 pav. Kalibravimo (kairėje) ir pavyzdžio (dešinėje) signalų Hilberto apgaubos, kai LFM sužadavimo trukmė yra 50 ms.....	130
130 pav. Nuokrypio paklaidos 10 ms 20 V LFM.....	131
131 pav. Laikinis (kairėje), spektrinis (viduryje) ir liekamasis (dešinėje) atsakas 50 μ s, kuris buvo mokymo dalis.....	132
132 pav. Laikinis (viršuje), spektrinis (centre) ir likutinis (dešinėje) 350 μ s atsakas, kuris buvo mokymo dalis.	132
133 pav. Nuokrypio NC-RUS paklaidos mėginio storio įvertinimui esant 20 V sužadimui (kairėje), 50 V sužadimui (viduryje) ir 100 V sužadimui (dešinėje).	134
134 pav. Atsitiktinės NC-RUS paklaidos mėginio storio įvertinimui esant 20 V sužadimui (kairėje), 50 V sužadimui (viduryje) ir 100 V sužadimui (dešinėje)	134
135 pav. Nuokrypio TRUS paklaidos, įvertinant mėginio storį (viršuje kairėje), greitį (viršuje dešinėje), tankį (apačioje kairėje) ir silpnėjimą (apačioje dešinėje)	135

LIST OF TABLES

Table 1. Resonance frequencies of materials with estimated thickness and velocities using 650 kHz air-coupled ultrasound simulation.....	27
Table 2. Total of attenuation in air at 20°C, 1 atm and 50% relative humidity.....	28
Table 3. Summary of air-coupled ultrasonic transducers, compiled from representative literature sources.	32
Table 4: SNR standard deviation with number of switch levels and band of frequency ranges.....	37
Table. 5. Remainder energy and correlation of each duration signals.....	78

LIST OF ABBREVIATIONS

ACQ – Acquisition control;
ADC – Analog to digital convertor;
APWP – Arbitrary pulse width and position;
AWGN – Additive white Gaussian noise;
BVD – Butterworth–Van Dyke;
BCZT – Barium calcium zirconate titanate;
BLE – Bluetooth low energy;
CRLB – Cramer-rao lower bound;
CMUT – Capacitive micromachined ultrasonic transducer;
dB – Decibel;
DC – Direct current;
DAC – Digital to analog converter;
DNN – Deep neural networks;
EMATs – Electromagnetic acoustic transducer;
FEM – Finite element method;
FM – Frequency modulated;
HV – High voltage;
IF – Inverse filter;
IIR – Infinite impulse response;
LReLU – Leaky rectified linear unit;
LFM – Linear frequency modulation;
 μ C – Microcontroller unit;
MF – Matched filter;
MMF – Mismatched filter;
NDT – Non-destructive testing;
NLFM – Non-linear frequency modulation;
NC-RUS – Non-contact resonant ultrasound spectroscopy;
NVSD – Noise voltage spectral density;
PMN-PT – Lead Magnesium Niobate–Lead Titanate;
PMS – Power management system;
PSO – particle swarm optimization;
PVDF – Polyvinylidene fluoride;
PZT – lead zirconate titanate;
PC – Polycarbonate;
RE – Reminder Energy;
ResUNet1D – Residual U-shaped network 1 dimension architecture;
RH – Relative humidity;
RUS – Resonant ultrasound spectroscopy;
SNR – Signal-to-noise ratio;
SS – Spread spectrum;
SOS – second order section;
ToF – Time-of-flight;
TRUS – Thickness resonance ultrasound spectroscopy;
 Q – Quality factor;
UNET – U-shaped network architecture;
WF – Wiener filter.

INTRODUCTION

Relevance of the research

Non-destructive testing (NDT) is essential for the safety, reliability and longevity of materials and structures across a multitude of applications [1,2]. Among the array of NDT techniques, ultrasound stands out for its ability to investigate in non-invasive manner [3]. Direct relationship between wave propagation and the material mechanical properties is the reason that ultrasonic measurements are so popular all surf [4], also it is a non-ionizing radiation, equipment is portable, low cost, low power. Ultrasound is widely used in biomedicine [5], aerospace [6], civil engineering [7], material testing [8,9], industrial processing [10], flow measurements [11], food industry [12,13], and even for sensitive material examination [14].

In general, ultrasonic measurement fundamentally involves the transduction of electrical energy into acoustic waves, coupling those waves into test sample and then conversion back to receiving electronics [1,2]. Analysis of the reflected or transmitted signals enables to characterize the material. The quality of such measurement depends on conversion efficiency of the transduction method. Ultrasound can be coupled into the sample using several mechanisms: direct acoustic contact, immersion, air-coupled, thermoacoustic, laser vibrometry, electromagnetic acoustic transducer (EMATs) [15,16].

In case of contact coupling, transducer and sample are in contact with each other [13]. Both surfaces exhibit some roughness, therefore transmission through contact is not repeatable, scanning is complicated. Therefore, coupling liquid (usually a thin layer of water, gel or oil) is used. Intermediate solution, dry coupling [17] is in between contact and immersion coupling. The main applications of the contact ultrasound are in the flaw detection, thickness gauging, weld inspection, adhesive inspection, erosion mapping, heat-affected zone evaluation, and inspection of high temperature components [2]. Contact ultrasound offers strong signal due to the minimal acoustic impedance mismatch; transducer can be quickly positioned on sample. However, fragile or rough surfaces might be unsuitable for direct contact and spatial resolution can be limited by the size of transducer and spread of beam in the inspection material, mechanical scanning requires constant supply of the coupling liquid, consistent coupling can be challenging [13]. Array transducers partly solve the aforementioned problems, but the aperture (so the resolution) is still limited by the array size. Yet, the challenge is to maintain the acoustic contact constant: surface roughness is changing, it is hard to maintain the right transducer inclination, surface abrasion destroys the matching layer thickness, integrity.

In case of immersion coupling the sample is submerged into the liquid medium (usually water). Immersion provides good coupling, highly focused transducers can be used, intricate geometry samples can be tested [18,19]. However, as sample needs to be in the liquid, such coupling is not suitable for fragile, water sensitive, or porous materials. Also, if equipment with multiple transducers is required (e.g. through-transmission) along with the requirement of elaborate scanning, transducers must be water-tight and as well as properly integrated into the positioning setup.

In air-coupled ultrasound, air itself serves as the coupling medium, enabling fully non-contact measurements [20]. Air-coupled ultrasound is ideal for porous, fibrous materials, food products, biological tissues, moisture sensitive and high temperature materials [21]. Scanning equipment does not have to be immune to water, complex scanning setups can be arranged, enabling quick measurement setup.

Laser ultrasound employs a pulsed laser to heat the sample surface which in turn generates the ultrasonic waves on material surface [22]. Laser vibrometry can be used to detect ultrasonic waves on the sample surface [23]. Usually, the Doppler shift of the laser beam reflected from the sample is used for mapping surface displacement with high spatial and temporal resolution. However, the use of laser (plasma regime uses $\geq 10^7 \text{ W/cm}^2$, 100 million times more than sun radiation) is related to eye (2.5 mW/cm^2) and skin damage (100 W/cm^2), laser ultrasound systems are bulky, expensive, require precise alignment, and can be sensitive to the surface conditions such as roughness and reflectivity, which may limit the use in some industrial applications [24]. Laser vibrometry requires a highly reflective surface, is subject to environmental vibrations, or ambient light interference.

EMATs generate and detect ultrasound in electrically conductive materials via electromagnetic induction, avoiding the need of coupling or physical contact [25]. This method has typically low energy conversion efficiency compared to piezoelectric transducers, which results in lower SNR. Also, crushing and pinching fingers or skin by strong magnets can occur [26]. Moreover, the inspection of EMATs is limited to conductive materials and may require specialized instruments for signal processing [27].

Only immersion [18], and mainly air [28] coupling is considered in this work. Air-coupled ultrasound is more affordable, safer in operation, offering portable and easy to operate equipment [29].

However, the acoustic impedance of air is about 410 Rayl, significantly lower than that of most solids (1-30 MRayl) [21,30]. As ultrasound waves traverse from air to solid material, most energy will be reflected at the interface, thus the transmitted signal is significantly reduced in round trip (air $\rightarrow$ solid $\rightarrow$ air). The loss for aluminium will be 80 dB [15]. Impedance mismatch results in extremely low amplitude signals and relatively low signal-to-noise ratio (SNR). Attenuation in the air is high and frequency dependent. At 100 kHz, the attenuation over 100 mm of air is relatively modest (about 1–2 dB), while at 1 MHz it is already 30–40 dB, and at 5 MHz it can exceed 200 dB [31]. Combined with the severe impedance mismatch at the air–solid interface—which can introduce an additional 60–90 dB of transmission loss, the total signal reduction for a round-trip (air $\rightarrow$ solid $\rightarrow$ air) can exceed 100 dB, especially at higher frequencies [5,31,32]. Therefore, air-coupled transducers are typically designed for low frequencies (10's of kHz to a few MHz) and operation is at tens of millimetres, where signal losses are more manageable.

There are two major operation modes for ultrasonic measurements: reflection and through-transmission. In reflection mode, the sample is evaluated just from one side, registering the waves, reflected from the inner structure of the material. The time-of-flight (ToF) of reflected longitudinal or Lamb wave [33], signal can be used to measure the depth of defect or material thickness (if velocity is known) or velocity

(from known thickness) [34]. The amplitude of reflection can be used to assess the impedance mismatch. Reflection mode, except the Lamb wave signals [33,72], is rarely used in air-coupled measurements, because the main part of energy comes from surface, masking the inner layers signals. Reflection mode is also associated with dead zone problem if same transducer is used for transmission and reception [35]. Through-transmission mode is using one transducer on one side of the sample to transmit the signal and another transducer on opposite side of the sample to receive [36]. Such a setup emphasizes the signals for sample inner structure therefore material properties can be studied. In its simplest form, signal amplitude can be used for internal defects evaluation [34]. Yet, ToF of multiple reflections inside the sample can be used to evaluate the material thickness (if velocity is known) or velocity (from known thickness).

In general, ToF is measured between two distinct signals. Then, the use of higher frequencies is desired to resolve two signals. It is the absolute bandwidth that matters in axial resolution, since a transducer's bandwidth is proportional to its centre frequency, using a higher frequency transducer provides a wider usable bandwidth. Higher frequencies are important for better lateral resolution.

Resonant ultrasound spectroscopy (RUS) is an alternative way of material mechanical properties evaluation. It relies on the natural vibration modes (eigenmodes) of the sample to determine its material mechanical properties [37–40]. RUS analyses the resonance response of the material. In a standard RUS setup, an excitation signal such as a swept sine wave or a broadband pulse is transmitted to the sample using a piezoelectric transducer or another suitable actuator. As the frequency of the input signal varies, the sample responds by vibrating at its natural resonance modes. The resulting mechanical vibrations are detected by a receiving transducer or sensor, which converts the response back into an electrical signal [37]. By systematically sweeping the excitation frequency across a desired range and recording the corresponding amplitude and/or phase of the received signal, a resonance spectrum is obtained. Peaks in the amplitude spectrum (and characteristic features in the phase spectrum) correspond to the material's resonance frequencies, which are in complex relationship to its elastic, geometric, and boundary properties. With geometry of the sample known, elastic properties matrix is obtained by solving the inverse problem through analytical expressions or finite element method (FEM) modelling [15,37].

While conventional RUS requires careful sample mounting to achieve free lightly constrained boundary conditions, thickness resonant ultrasound spectroscopy (TRUS) is using air [21,29] or water [18] coupling media which is much easier to arrange technologically. Usually, two measurements are carried out for plate parameters evaluation: i) Calibration measurement, when path between transducers is free from obstacles and ii) Sample measurement, when sample is inserted between transducers. The plate's thickness resonances manifest as multiple peaks in the transmission spectrum, and these peaks are determined by the sample thickness and velocity. Also, there is a signal shift in time between Calibration and Sample measurements which is also related to sample thickness and ultrasound velocity. Such two independent measurements allow to simultaneously estimate the sample

thickness, density, ultrasound velocity and attenuation by arriving to inverse solution by sample model. TRUS was designed for cases when bandwidth is so narrow that the reflections inside the material overlap and the traditional RUS or ultrasound techniques cannot be used. This ability allows TRUS a powerful inspection method for the characterization of small and complex objects, including plant leaves [32,36,41], where contact or immersion methods fail. In the case of air-coupled TRUS such measurement is truly non-contact, therefore sometimes it is addressed as non-contact resonant ultrasound spectroscopy (NC-RUS) [36,42]. Detailed explanation of TRUS technique is presented in section 2.2. The focus of this study is to improve the TRUS measurements, the target objects are thin plates and plant leaves and the primary metrological parameters are: sample thickness (h), density (ρ), ultrasound velocity (c) and attenuation (α).

The impedance mismatch between coupling media and sample results in significant reflection losses at the interface [43]. Thus, the energy transferred through the sample is less, lowering SNR and susceptibility to electromagnetic interference. Attenuation depends on the frequency in both the coupling medium (such as air or water) and the test sample. The effect is especially pronounced in air, where attenuation increases dramatically with frequency, reducing SNR further and limiting the practical use of high-frequency ultrasound in air-coupled applications. Therefore, higher ultrasonic frequencies experience even greater attenuation, which in turn limits the effective bandwidth available for inspection, particularly in air coupled [43]. Usually [21,29], a short duration ultrasonic pulse signal is used for excitation to get a broadband excitation [44]. For pulse, energy and bandwidth contradict each other, so with shorter pulse SNR is further reduced. With single short-pulse excitation, the spectral shape is defined by the sinc function and cannot be freely controlled. As a result, the energy distribution across frequencies cannot be selectively shaped, and additional energy cannot be directed into specific frequency band.

Spread-spectrum (SS) signals are not constrained by the fixed duration–bandwidth relationship of short pulses and can therefore deliver significantly higher SNR than single-pulse excitation. The considered SS signals include chirp waveforms (linear frequency modulation, LFM) [45,46], non-linear chirps (NLFM) [47,48], and arbitrary pulse-width and position (APWP) waveforms [36,42,49–51]. Spectral coverage of SS signals is defined by spreading parameters. SS are used for their ability to spread energy in time, moreover, SS can be programmed for spectral shaping, therefore, energy can be pushed into the frequencies where losses are higher, which broadens the transmission spectrum (increased bandwidth and reduced Q-factor) [49,50,52].

The last challenge lies in SS signal itself. The spectral shape flexibility of APWP signal depends on signal length: the longer is the signal, the more flexible and more accurate is the spectral shape. Long SS signal duration is important to accumulate sufficient energy for achieving broad bandwidth: it requires distributing energy away from the transducer's centre frequency, where transduction efficiency drops significantly. However, in TRUS the transducers must be placed relatively close to the sample in order to operate within the near field, maintain a sufficiently flat wavefront, and minimize air attenuation. Because both the sample surface and the

transducer face have high acoustic reflectivity, multiple reflections occur between them. These repeated reflections overlap with the first transmitted signal, complicating the interpretation of the measured waveform and degrading the accuracy of the inverse solution [49].

Solution seems trivial: SS signals are compressible. Pulse compression can be achieved by matched filter (cross-correlation processing). Yet, transducer response has a long ringdown due to multiple matching layers. Therefore, the level of correlation sidelobes is high. Signals after such pulse compression cannot be gated. It was proposed to approach the problem from two directions: i) use of some derivative of inverse filter and ii) deconvolution of multiple reflections. The inverse filter proposed is derived using the excitation gencode (generation code) or signal on transmitting transducer clamps. Inverse filters possess the same phase response as matched filter, so signal compression occurs. But the result of such filter application is not a matched filter signal, that has symmetrical sidelobes but is equivalent to delta-pulse excitation. Better or equivalent to pulse excitation results are achieved. Then, signal can be either gated with same success as single pulse signal or fitting of inverse solution performed in time domain. Results were compared against matched, Wiener and mismatched filter processing.

Finally, single transducer pair measurement may become feasible with sufficient spectral coverage and high SNR without averaging. Ability of SS signals to provide higher SNR, together with development of new highly efficient air-coupled transducers [21,29,32,43], allows to consider NC-RUS as a reliable and operative tool for material properties evaluation.

Hypothesis: Quality of the resonant ultrasound spectroscopy measurements can be improved if long spread spectrum signals used for excitation have spectral shape programmed to compensate for the spectral losses.

Aim of the research: to develop a method for spectral losses compensation using programmable spectral content signals and investigate the performance in resonant ultrasound spectroscopy measurements.

Research tasks

1. To develop a technique for compensating transduction spectral losses and evaluate its effectiveness in improving TRUS measurements using a single transducer pair and non-overlapping signals.

2. To develop optimization criteria for programmable spectral shaping and determine whether transduction losses compensation alone suffices for TRUS improvements or if the sample's frequency response must also be considered.

3. To design and investigate techniques to mitigate the effects of the long spread spectrum signal overlapping in TRUS measurements.

4. To evaluate the effectiveness of spectral losses compensation when using long duration excitation signals in TRUS.

Scientific Novelty

1. Previous studies have primarily focused on optimizing the excitation signal spectrum for loss compensation by programming the spectral shape [49,51,53]. However, the frequency-dependent nature of noise was not accounted. In this work, optimization is based on the SNR, recognizing that effective compensation requires accountability of both the signal and the noise characteristics across the frequency range.

2. It was demonstrated that for spectral compensation bipolar excitation signal can be replaced with unipolar, which significantly reduces the complexity of the required electronics.

3. A generalized spectral optimization framework for ultrasonic signals was proposed, enabling the design of arbitrary spectral shapes tailored for signal spectrum.

4. A minimum spectral coverage criterion for TRUS was identified: the excitation bandwidth must include at least one resonance peak and its adjacent valley for stable and low-bias parameter estimation under the investigated conditions.

5. It was demonstrated that inverse filtering-based pulse compression provides signal equivalent to delta-like excitation, enabling long signals compression and gating. This technique is expandable to other types of coupling and other types of ultrasonic measurements and imaging.

6. Residual deep learning model was proposed for long duration overlapping signals.

Practical value of work

The developed methods enable non-contact measurement of mechanical parameters in thin, fragile, or moisture-sensitive samples, such as biological tissues, plant leaves, thermal insulation materials, paper, cardboard, filters, membranes, or unpolymerized composites. By combining air-coupled ultrasound with through-transmission TRUS and innovative signal processing—including long programmable spectrum signals, pulse compression, and spectral losses compensation—the approach improves measurement quality using a single transducer pair, overcoming high-frequency attenuation challenges.

The pulse compression technique yields responses equivalent to high-energy broadband pulses with minimal artifacts, enhancing SNR and resolution. This extends beyond TRUS to any ultrasonic application requiring high SNR and precision, including immersion or contact measurements.

Furthermore, advanced signal decomposition and deep learning-based reflection separation facilitate the use of long-duration pulses in reflective or layered media, preserving spectral resolution for accurate material property inference.

Finally, TRUS can be scaled to 2D or 3D imaging of properties like density, elastic modulus, and thickness via spatial scanning and focused beams, supporting applications in industrial inspection, biomedical diagnostics, and agricultural quality control.

Results of the research were applied to the following projects:

- New non-contact ultrasound transduction methods for wideband spread spectrum signals” from Research Council of Lithuania, Project no.: S-MIP-23-133 (2023-07-01 - 2025-10-31)

- Ultrasound spectroscopy sensor for plant physiology studies” by the European regional development fund, Project no.: 01.2.2-LMT-K-718-03-0026 (2020-07-20 - 2023-08-31)

Scientific approbation

The scientific results obtained during the course of dissertation were presented as 10 scholarly contributions: 3 articles were published in international journals referred in the Web of Science DB (Q1), results were presented in 8 international conference presentations (3 poster and 5 oral) in Lisbon (Portugal), Athens (Greece), Taipei (Taiwan), Kaunas, (Lithuania), Paderborn (Germany) and Vilnius (Lithuania). 4 conference proceeding papers were published.

Results presented in the defence of thesis

1. A generalized APWP optimization framework was developed and experimentally validated. It enables spectral programming according to measured SNR profiles, transducer responses, or target resonances, resulting in significantly improved signal quality: the -20 dB usable bandwidth increased from approximately 480 kHz (conventional pulse and chirp signals) to 670 kHz.

2. Application of the optimized APWP signals in air-coupled TRUS were shown to reduce measurement errors while allowing the use of only a single transducer pair across a wide frequency range, overcoming the previous need for multiple transducer sets in air-coupled ultrasonic measurements.

3. Solution for overlapping reflections of long spread spectrum signals by using inverse filter compression, gating and decompression was developed and validated. Moreover, deep learning framework for reflection modelling was developed, achieving ~ 45 dB accuracy in reminder energy (RE) and high correlation (~ 0.9) for 50 μ s duration and for 300 μ s RE: ~ 32 dB, correlation: ~ 0.74 between predicted and experimental signals.

4. It was demonstrated that spectral losses compensation using long APWP signals improves the overall performance of TRUS measurements in highly attenuating air-coupled conditions, while accurate resonance parameter estimation requires the excitation spectrum to cover at least one full resonance peak together with an adjacent valley.

Structure and coverage of the thesis

The dissertation consists of an introduction, five chapters, conclusions, summary in Lithuanian language and references. The volume of the dissertation is 154 pages, with 5 tables, 53 equations, 117 figures and 143 references and summary of research in Lithuanian language. The content of thesis is as follows:

The first chapter discusses the air-coupled RUS, its applications and importance for NDT measurements

The second chapter presents the development of SNR based APWP signals using spectral programming and compensation function development for NC-RUS, with performance evaluation.

The third chapter is used to assess the influence of the spectral content of the signal on the quality of thickness resonance spectroscopy measurements. It is investigated how the errors of the inverse solution change by changing the frequency range in which spectral losses are compensated with respect to resonant maxima and minima. It is also investigated how various proposed spectral content optimization criteria affect the quality of the inverse solution.

The fourth chapter presents the development of inverse filter-based pulse compression and residual deep learning methods for solving the overlapping of long spread spectrum signals.

The fifth chapter provides inverse filter-based pulse compression application performance evaluation in air-coupled and immersion TRUS using long spread spectrum signals.

1. PROBLEM ANALYSIS

1.1. Time of flight estimation

Accuracy of ultrasonic measurements (distance, ToF, imaging etc.) depends on SNR and sharpness (effective bandwidth) of the signal [60,61], often expressed as Cramer-Rao lower bound (CRLB), which shows that timing variance decreases with higher SNR and larger signal bandwidth. The focus here is on thickness measurements.

Ultrasonic transducers produce an acoustic pulse that propagate through the material. As there are multiple interfaces or layers inside the material, at each layer there is a change in acoustic impedance. As a result, a portion of wave is reflected and returns to the transducer, while remainder is transmitted [58]. The received signal is therefore sum of time-delayed, scaled and possibly, overlapped versions of the transmitted pulse superimposed with measured pulse [59,61].

Fig. 1 demonstrates the case when second reflections start to overlap with first reflection and are not distinguishable.

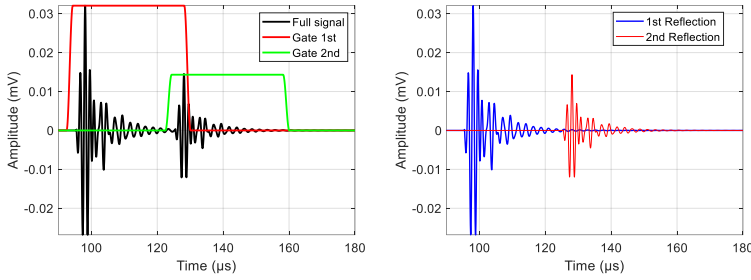


Fig. 1 (left) Partially overlapping reflections and (right) isolated plot of the 1st and 2nd reflections, shown separately for clarity

Furthermore, in case of severe overlapping or complete overlapping of reflections (Fig. 2) the individual reflections overlap, especially if the transmitted pulse is wide but not short enough in duration. In such cases, the measured signal may appear as a single broadened or distorted peak, making direct ToF extraction impossible. Furthermore, experimental noise masks these features, often producing misleading or spurious peaks.

Deconvolution is a prominent approach providing the impulse response ($h(t)$), from the measured ($S_R(t)$), response if excitation or reference pulse ($S_T(t)$), is known. The frequency domain representation of deconvolution theorem is as,

$$S_R(f) = S_T(f) \cdot H(f) + N(f), \quad (1.1)$$

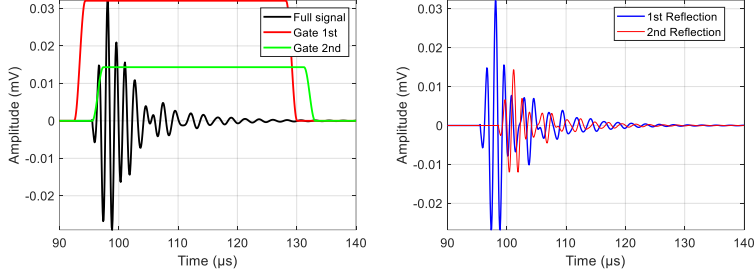


Fig. 2. (left) Fully overlapping reflections and (right) isolated plot of the 1st and 2nd reflections

Simple division amplifies noise N at frequencies where S_T is small, leading to unstable results. The Wiener deconvolution approach regularizes this inversion,

$$H(f) = \frac{S_T^*(f)}{|S_T(f)|^2 + \kappa} S_R(f), \quad (1.2)$$

where, κ is the regularization parameter, and S_T^* is the complex conjugate of the excitation or reference spectrum. Wiener deconvolution provides a tradeoff between noise amplification and resolution, and the value of regularization parameter is often selected empirically.

The Wiener deconvolution for the overlapping signal can be seen in Fig. 3, by dividing signal spectrum with spectrum of the reference (excitation) signal. This should compress the reflections into two sharp impulses at the reflection locations, provided SNR is high.

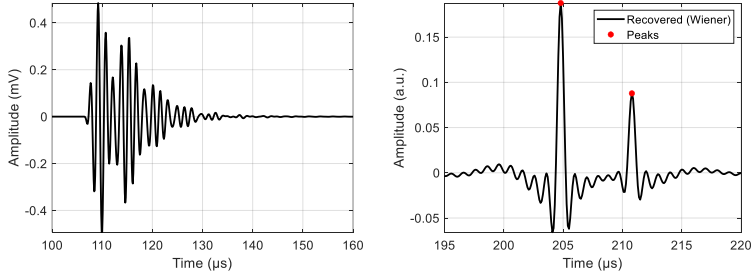


Fig. 3. (left) Time domain overlapping of reflections for noise free signal, (right) Wiener deconvolution (regularization parameter = 10^{-12}) when reflections are overlapping for noise free signal

Although Wiener deconvolution is a promising approach [91], it certainly has limitations: as noise level increases it disrupts the deconvolution process.

1.2. Thickness resonant ultrasound spectroscopy

RUS is classical vibrational analysis method, focusing on the natural resonance frequencies within the sample to determine the fundamental properties [40,62]. Based

on elasticity, geometry, and density, every material vibrates at a set of natural frequencies (resonant frequencies) when excited [38,63]. Classic RUS investigation involves measuring many resonance frequencies of sample and then solving inverse problems to find elastic constants. RUS principles were used for the development of TRUS [42,50,64], where primary focus is to estimate the thickness (h), density (ρ), velocity(c) and attenuation (α) parameters of sample simultaneously using inverse solution of TRUS. The fundamental thickness resonance of sample in air occurs when the half-wavelength equals the thickness:

$$f_n \approx \frac{n \cdot c_L}{2 \cdot h}, \quad (1.3)$$

where, c_L is the longitudinal sound velocity in the sample and h is the thickness, higher harmonics correspond to the $n = 1, 2, \dots$ wavelengths. The resonance spectrum is sensitive to the sample parameters (h, ρ, c, α), as illustrated in Fig. 4.

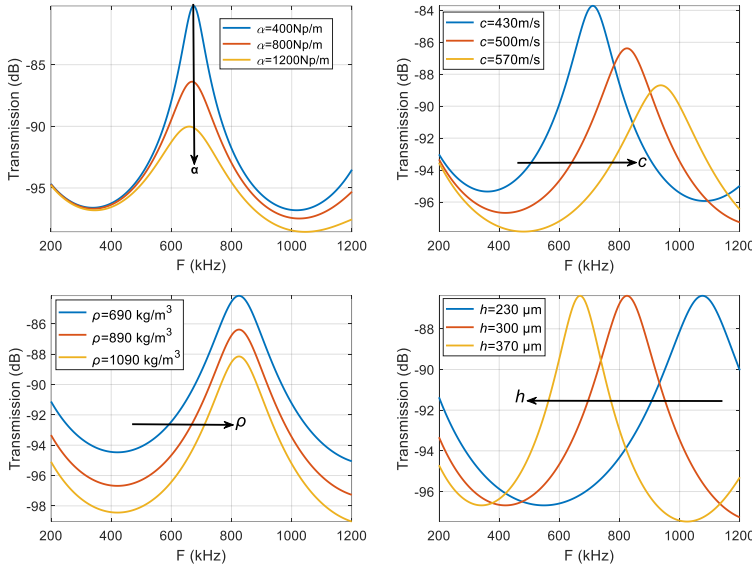


Fig. 4. Effect on transmission response of (top left) sample attenuation, (top right) velocity, (bottom left) density and (bottom right) thickness

Increasing sample attenuation broadens the resonance peak, indicating stronger energy losses in the sample (Fig. 4, top left). Increase of sample velocity c shifts the resonance peak towards higher frequencies confirming the relationship in Eq. 1.3, moreover, the spacing between the resonances are also increased (Fig. 4, top right). Increase in density results in the decrease of resonance amplitude due to acoustic impedance mismatch, the change in the resonance frequency is minimum (Fig. 4, bottom left). Lastly, increase in sample thickness h , shifts the resonance towards lower frequencies, consistent with relationship $f \propto 1/h$. Thus, thinner samples resonate at higher frequencies (Fig. 4, bottom right). The resonance peak position is most sensitive to velocity and thickness, while the peak amplitude and width depend strongly on density and attenuation, respectively. In practice, this constitutes a well-

posed inverse problem: by fitting simulated spectra to experimental data, all four parameters (c , h , ρ , α) can be estimated simultaneously. The resonance frequency of various materials is shown in Table 1.

Table 1. Resonance frequencies of materials with estimated thickness and velocities using 650 kHz air-coupled ultrasound simulation

Note: although the physical transducer bandwidth is centred at 650 kHz, a broadband transducer model (obtained through infinite impulse response filter) was used to simulate the material transmission response; full details are provided in Chapter 2.

Material	$h(\mu\text{m})$	$c_L(\text{m/s})$	$\alpha(\text{Np/m})$	$\rho(\text{kg/m}^3)$	$f_1(\text{MHz})$	$f_2(\text{MHz})$
Leaf (Annona Cherimola)	85.20	173.0	599.7	1548	1.015	2.031
Leaf (Alocasia)	589.2	251.3	3107	427.9	0.213	0.426
Leaf (Amstel Queen)	699.9	190.4	2825	1277	0.136	0.272
Leaf (Dwarf Cavendish)	263.0	368.0	1717	1429	0.754	1.399
Leaf (Grape leaf)	250.1	315.3	800.2	840.8	0.798	1.597
Leaf (Persea Americana)	159.6	252.7	2276	890.8	0.792	1.583
Leaf (Ctenanthe Setosa)	217.1	412.0	1316	740.2	0.949	1.897
Card paper	270.5	290.1	1650	870.1	0.950	1.901
Steel	4538	5900	2.100	7850	0.650	1.300
Polycarbonate	2045	2240	43.5	1193.6	0.548	1.095
Thick card paper	250.0	184.0	3106	2387	0.368	0.736
Plexiglass	1015	2720	9.200	1180	1.340	2.680
Aluminium	2500	6350	0.200	2700	1.270	2.540

Resonance frequency mainly depends on the thickness and velocity of the sample; transmission response depends also on the attenuation and density of the sample. The transmission response of different samples organic (plant leaves) and non-organic (paper, polycarbonate, plexiglass, aluminium, steel) are shown in Fig. 5.

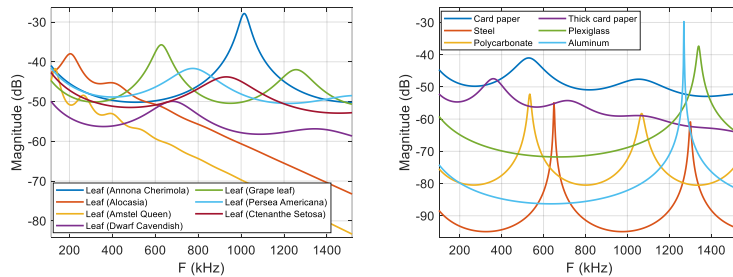


Fig. 5. Transmission response of (left) organic and (right) non-organic samples

Parameter estimation is then performed via optimization algorithms, using the sensitivity analysis it can determine which parameters are more strongly constrained for spectral features [36,42].

From Table 1 and Fig. 5 it can be deduced that position of the resonant peak can vary widely over frequency range, depending on sample parameters. It can be noted that many samples have a peak in range from 150 kHz to 1.5 MHz. Thicker samples can have lower than 150 kHz fundamental resonance peak, but TRUS can work on higher order resonance peak. Also, from Fig. 5 it can be noted that attenuation in sample propagation is from -30 dB to -100 dB. Then it can be concluded that TRUS is demanding wide bandwidth and high SNR.

1.3. Acoustic wave propagation

Acoustic wave propagation in air or medium relies on many interrelated phenomena that affect the amplitude, spatial distribution and speed of wave. The three key effects are attenuation, and speed of sound variation. Moreover, the critical angle near field distance are also important features of wave propagation that must be considered while developing new approach.

Sound propagation in air is a complex phenomenon as energy losses occur due to the interaction of sound waves with air molecules [65]. There are many mechanisms that influence the attenuation: classical (viscosity, thermal conduction), rotational relaxation, and vibrational relaxation of oxygen (O_2) and nitrogen (N_2) molecules. These mechanisms are affected by environmental parameters (temperature, atmospheric pressure, relative humidity (RH) and sound frequency [65].

Attenuation frequency dependence is shown in Fig. 6.

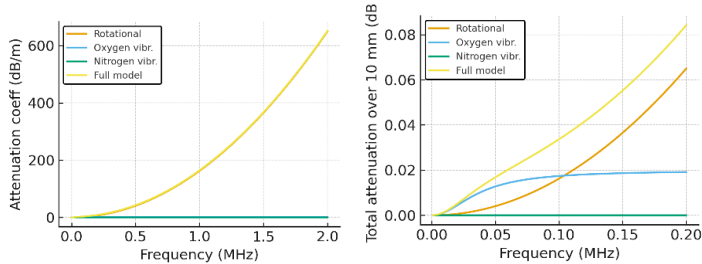


Fig. 6. Frequency-dependent sound attenuation in air at 20°C, 1 atm, with 50% relative humidity and 10 mm distance path, (left) attenuation over 2 MHz frequency range, (right) attenuation over 200 kHz frequency range

Total attenuation over specific distance at different frequencies are presented in Table 2.

Table 2. Total of attenuation in air at 20°C, 1 atm and 50% relative humidity

Frequency	10 mm	20 mm	50 mm	100 mm
100 kHz	0.034 dB	0.067 dB	0.168 dB	0.337 dB
200 kHz	0.084 dB	0.168 dB	0.421 dB	0.841 dB

500 kHz	0.426 dB	0.851 dB	2.129 dB	4.257 dB
1 MHz	1.644 dB	3.288 dB	8.219 dB	16.43 dB
2 MHz	6.516 dB	13.03 dB	32.57 dB	65.15 dB
5 MHz	40.62 dB	81.24 dB	203.1 dB	406.1 dB

It can be noted that sound attenuates only 0.034 dB over 10 mm distance at 100 kHz is, while attenuation at 10 mm for 5 MHz is already 40.62 dB, and at 100 mm and 5 MHz is huge, 406.1 dB. Therefore, long range operation is possible only below 500 kHz, operation beyond 2 MHz should be limited to 100 mm.

The speed of sound in an ideal gas like air is fundamentally determined by the gas's thermodynamic properties. An ideal diatomic gas speed of sound c is described as [68]:

$$c = \sqrt{\gamma \cdot R \cdot T}, \quad (1.4)$$

where, γ is the ratio of specific heats (approximately 1.4 for air at moderate conditions), R is the specific gas constant for air and T is the absolute temperature [69,70].

The influence of pressure on c_0 (Fig. 7, left), where speed of sound is decreasing nonlinearly with pressure from ~ 343.15 m/s at 1 kPa, to ~ 343.05 m/s at 80 kPa, which corresponds to the altitude of ~ 2 km. The effect of pressure stems from slight changes in molar mass and γ , as pressure increases, the speed of sound slightly decreases. The results conclude that, at ambient pressures (80–120 kPa), changes in pressure barely affect sound speed, that's why speed of sound in air is typically assumed constant at 1 atm (≈ 101.325 kPa).

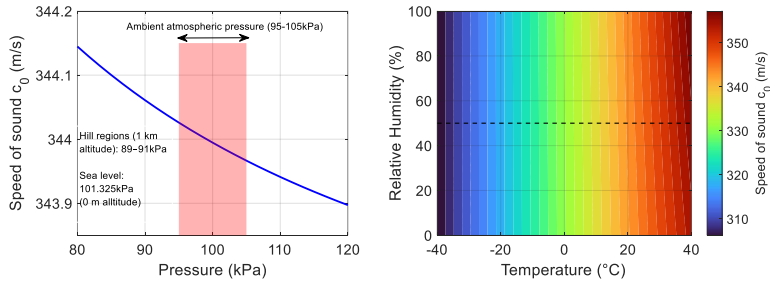


Fig. 7. (left) Atmospheric pressure vs. (right) temperature and relative humidity influence on speed of sound

Speed of sound dependency on temperature (Fig. 7, right), across -40 to 40°C and 0 to 100% humidity. Resulting range of speed of sound is from 310 m/s (cold, dry) to 360 m/s (warm, humid). The essential influence is from temperature, and humidity is adding ~ 0.5 m/s variations per 100% . Temperature is the dominant factor affecting sound speed in air. Even a 10°C change causes ≈ 6 m/s difference.

The results demonstrate that the speed of sound in air is not constant but varies with temperature, humidity, pressure, and, to a lesser extent, frequency. These dependencies are critical in ultrasonic measurements, where the sound speed directly determines wavelength, phase velocity, and resonance conditions. Even small

variations can lead to noticeable shifts in resonance frequencies and apparent material thickness when performing air-coupled resonant spectroscopy. Therefore, accurate modelling and real-time correction of the sound speed under actual ambient conditions are essential to ensure precise frequency–thickness calibration, reliable estimation of acoustic impedance, and repeatable spectroscopic characterization of materials. Publication [43] developed a method that compensates for the speed of sound in, which will be used in further processing of the proposed work.

Critical angle [71] is important in evaluating which part of the beam can enter the sample. For $\theta > \theta_{crit}$, total reflection occurs at the interface (no transmitted longitudinal wave). The critical angle is calculated as

$$\theta_{crit} = \sin^{-1} \left(\frac{c_{air}}{c_{sample}} \right), \quad (1.5)$$

where, c_{air} is the speed of sound in air (e.g. 342 m/s, air at room temperature) and c_{sample} is the speed of sound in sample (e.g. 6400 m/s for aluminium). The critical angle for longitudinal wave is 3.1° for air-aluminium interface, 7.3° for air-plexiglass interface and 9° for air-polycarbonate interface [73]. The shape of the beam is illustrated in Fig. 8, where it can be seen that wavefront changes its shape from flat to spherical once it move away from its starting point (transducer). In must be pointed out that spherical waveform is not able to enter the sample completely, only part of the wavefront that hits the surface at less than critical angle will penetrate.

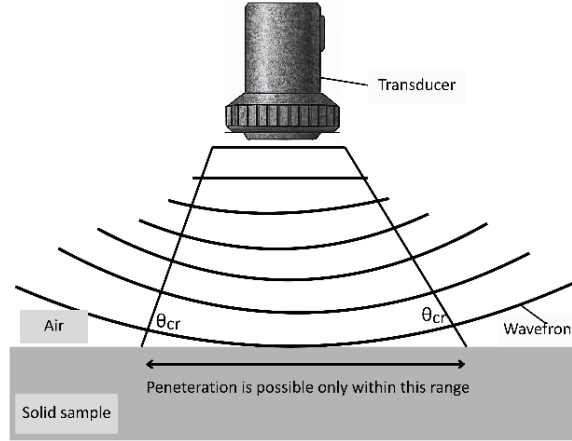


Fig. 8. Air-solid interface using air-coupled ultrasonic transducer

It can be concluded that flat beam is preferred in air-coupled ultrasonics and sample should be placed at less than near field distance in order to keep the beam flat [88].

Ultrasound beam pattern depends on the transducer diameter (D), excitation frequency (f), and medium speed of sound (c) [74,75]. Ultrasound propagation fields are typically divided into two zones: the near field (Fresnel zone) and the far field (Fraunhofer zone). The classical near-field N_f formula for a circular piston transducer with aperture D , wavelength λ [138, 75] is:

$$N_f = \frac{D^2 - \lambda}{4 \cdot \lambda} \approx \frac{D^2}{4 \cdot \lambda}. \quad (1.6)$$

The near field distance for 20 mm diameter transducer at 2 MHz is 500 mm, but is just 1.8 mm for 6 mm transducer at 100 kHz. Operation at half of such distance is preferred in order to keep the wavefront is significantly flat. It can be concluded that if transducer is operating from 100 kHz then its aperture should be at least 20 mm and operated at less than 30 mm distance in order to keep the beam flat.

1.4. Ultrasonic transducers

Transducers are usually based on piezoelectric effect. The most used piezoelectric materials are polarized ferroelectric ceramics such as lead zirconate titanate (PZT), polymer piezoelectric (Polyvinylidene fluoride (PVDF)) or single crystal (Lead Magnesium Niobate–Lead Titanate (PMN-PT)) materials. Since the acoustic impedance of a PZT ceramics is about 30–40 MRayl, compared to 1.5 MRayl for water and 0.0004 MRayl for air, there is a significant impedance mismatch. The efficiency of the transducer can be increased by adding a quarter wavelength layer with an intermediate impedance, which allows for the transmission of much more energy into the medium. Also, a well-designed matching layer can increase the transducer's bandwidth.

The cross-sectional schematic of typical transducer is shown in Fig. 9, the core is active piezoelectric element, coated on both faces with conducting electrodes.

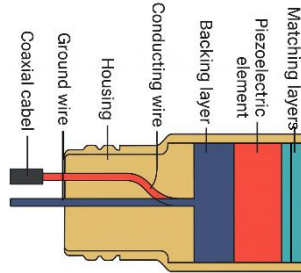


Fig. 9. Schematic of transducer

The resonance of the piezoelectric elements is mainly determined by its thickness, in the simplest thickness-extensional mode, the centre frequency (f_0) is approximately set so that thickness of the piezoelectric plate is one half of the acoustic wavelength inside the piezo material. A transducer intended for 5 MHz operation might use a PZT disk only around 0.4 mm thick (given typical in PZT sound speed ~ 4000 m/s), while a 500 kHz transducer's crystal could be on the order of several millimetres thick. Together with significant impedance mismatch this makes the emission very narrowband. Resonance quality can be reduced by adding backing, but this only reduces transduction sensitivity.

Typically, two or more quarter-wavelength ($\lambda/4$) matching layers are used in air-coupled transducers, but bandwidth limitation still exists [119].

Analysis of the literature on the development of air-coupled ultrasonic transducers reveals that piezoelectric material-based transducers can reach 65% bandwidth [67], while capacitive transducers can reach 140% bandwidth, yet their efficiency is much lower. Measured transmission response of $\lambda/4$ matched PZT 650 kHz centre frequency air-coupled transducer (Fig. 10, left). Same type, but 1 MHz centre frequency transducer transmission response (Fig. 10, centre). Capacitive transducer response (Fig. 10, right). The transmission bandwidths at -6 dB, -10 dB, -15 dB, -20 dB, -30 dB, and -35 dB are shown for both 650 kHz and 1 MHz cases.

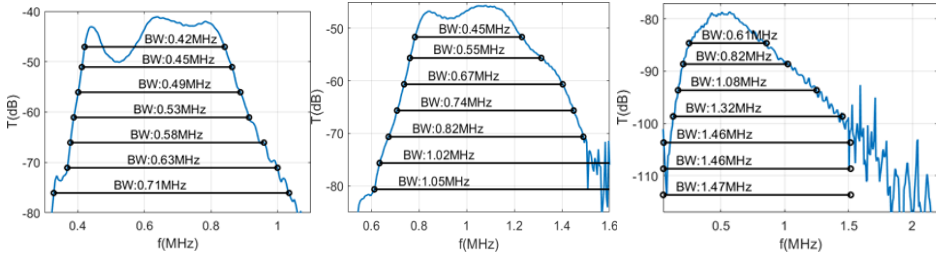


Fig. 10. Transmission response of $\lambda/4$ matched PZT (left) 650 kHz centre frequency transducer, (middle) 1 MHz centre frequency transducer and (right) capacitive transducer pair [140]

The bandwidth is broader for 1 MHz, 0.74 MHz at -20 dB versus 0.53 MHz for 650 kHz. The -30 dB and -35 dB bandwidth is broader, but compensation can become impossible due to large compensation depth required. Capacitive transducer -20 dB bandwidth reaches 1.32 MHz, nearly twice that of PZT transducers, but capacitive transducers exhibit lower sensitivity (-80 dB peak transmission compared with -40 dB and -45 dB for PZT), making TRUS measurements complicated.

Table 3 summarizes representative strategies and material innovations in the development of air-coupled ultrasonic transducers. Reported frequencies, bandwidths, and sensitivities are extracted/estimated from experimental data or graphs in the cited works.

Table 3. Summary of air-coupled ultrasonic transducers, compiled from representative literature sources

Sources	Material & Design	Centre Frequency (MHz)	Bandwidth (-6 dB)	Sensitivity
Hutchins et al., 2002 [76]	Capacitive air-coupled transducer (focused)	0.4	$\approx 175\%$	≈ -35 dBV (~ 50 mV pk-pk)
Song & Chimenti, 2006 [77]	Capacitive air-coupled transducer (focused), with aluminized Mylar foil backplate.	0.84	$\approx 99\%$	≈ -40 dBV

Satyanarayan et al., 2010 [78]	Cellular polypropylene foam (ferroelectret), with Plexiglass backing	0.33	$\approx 40\%$	≈ -50 dBV
O’Leary & Harvey, 2014 [79]	Capacitive micromachined (polymer-based CMUT), with Polystyrene backing	0.5	$> 100\%$	-62 dB (two-way insertion loss)
Nagahara et al., 2005 [80]	PZT piezoelectric ceramic, with Two-layer composite and metal case + epoxy bond backing	0.5	$\approx 15\text{--}20\%$	≈ -40 dBV
Huang et al., 2006 [81]	Capacitive micromachined (CMUT), with Silicon substrate bonded backplate	2.3 MHz	126–140 %	≈ -52 dBV (24 mV/kPa)
Wu et al., 2021 [82]	PZT/epoxy 1–3 composite (piezoelectric), with single closed-cell polymer composite	0.52	23%	≈ -33 dB
Zhou et al., 2024 [83]	Lead-free piezoceramic (BCZT), with hollow glass epoxy composite	0.54	7.4%	Insertion loss (two-way) ≈ -70 dB
Alvarez-Arenas, 2004 [102]	PZT-5A ceramic, with Epoxy + porous membrane and air backed	0.45 – 2.25	15%– 20%	-24 dB to -50 dB
Alvarez-Arenas & Diez, 2013 [104]	Piezoelectric composite, with 3 rigid polymer-foam layers	~ 1.0	55-65%	$-38\text{--}42$ dB

Sensitivity is expressed in dBV, which represents the transducer’s output voltage level in decibels relative to 1 V “RMS”. Because many papers report signal amplitude only in peak-to-peak or millivolt units, the values were converted to RMS and then to dBV for consistent comparison across the literature

1.5. Excitation signals for ultrasonic measurements

The selection of excitation signals in ultrasonic measurements involves a trade-off between bandwidth (for resolution), energy (for penetration and SNR), and system constraints, tailored to the specific measurement requirements such as ToF accuracy or RUS. The most used excitation signals for ultrasonic measurements are short pulse, spike or toneburst, as they are preferred for simplified generation electronics and easy interpretation [84,85]. However, such signals have fundamental limitations, as short pulse and spike have broad bandwidth but low energy, while longer toneburst increases energy at the expense of bandwidth [86,87]. Thus, these signals cannot simultaneously provide high energy and broad bandwidth [84,87]. Compared to simple pulse, spike or toneburst excitation signals, spread spectrum and arbitrary waveforms offer wider bandwidth, however they are quite complex to generate and difficult to interpret.

The key limitation for short pulse is low energy, for the fixed amplitude, the energy E is proportional to the duration (or number of cycles). A multi-cycle toneburst thus delivers more energy into the material, which can improve reflection strength and SNR. Indeed, increasing the toneburst length increases the SNR in proportion to the pulse duration (since noise is approximately constant over the short measurement window) [86]. However, simply lengthening a pulse without changing its bandwidth does not improve the time resolution, as long unmodulated signals cannot be compressed in time via processing.

SS signals do not have bandwidth and duration relationship, therefore can deliver higher energy without sacrificing the bandwidth. The most known SS signals are frequency modulated (FM) chirp signals (linear or non-linear FM sweeps) and phase shift keying (binary pseudorandom codes such as Barker codes, Golay complementary pairs, or m-sequences).

The linear chirp could achieve frequency sweep from frequency f_1 to f_2 over duration T_c , therefore covering the effective bandwidth $B = f_2 - f_1$, with $TB = B \cdot T_c \gg 1$. This type of chirp is useful for encoding and can be effectively used for matched filtering via compression which results in sharp peak. Linear frequency modulation (LFM s_L) is described as:

$$s_L(t)|_{t=0 \dots T_c} = A_L \cdot \sin\left(f_1 \cdot t + \frac{DF}{2 \cdot T_c} t^2\right), \quad (1.7)$$

where, DF is the frequency sweep (or deviation), A_L is the amplitude of LFM.

The non-linear frequency modulation (NLFM) s_N is described as:

$$s_N(t)|_{t=0 \dots T_c} = A_N \cdot \sin(f_N(t)), \quad (1.8)$$

where, f_N is the instantaneous phase, which can be any non-linear function.

The chirp signals offer several advantages: compression, analytical description and ease of generation. The spectral content of the chirp signals depends on f_1 and DF , so the width of correlation main lobe can be controlled. Similarly, the amount of delivered energy can also be controlled through A or T_c . NLFM offers programmable spectral shape, avoiding spectral zeros, which are not desired in TRUS and inverse problem [41,91].

The use of chirp signal is limited by their disadvantages, the constant envelope: convolution of spectral response with *sinc* function results in envelope spectra due to sharp rise and fall of envelope. This introduces Fresnel ripples in far range sidelobes, while rectangular shape of chirp produces near range sidelobes. The temporal response of LFM for 50 μ s with $f_1=350$ kHz and $f_2=950$ kHz is shown in Fig. 11 (left), and the spectral response is shown in Fig. 11 (right).

The correlation function of chirp results in *sinc* like envelope which does not depend on the time-bandwidth product. Tapering is an approach in which signal is multiplied with smooth window (Tukey, Hanning, etc.). Although the impact of tapering is not directly visible in spectral response, the autocorrelation results (Fig. 12), clearly demonstrate the suppression of sidelobes.

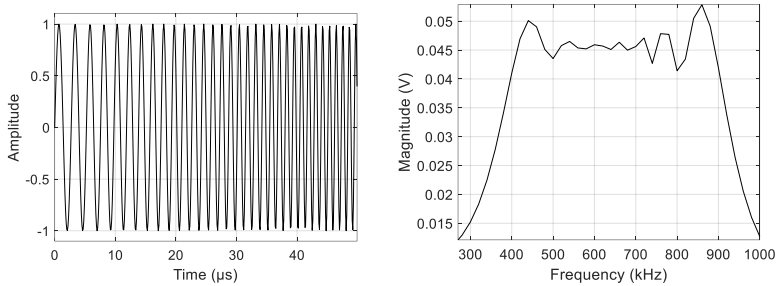


Fig. 11. (left) Sinusoidal chirp LFM in time and (right) frequency domain

Less known SS signal are arbitrary pulse width and position (APWP) waveforms [36,42,49–51]. APWP spectral content can be programmed to compensate the frequencies where losses are higher and thus response of the received signal can be made broader [49,50,52]. One more challenge lies in SS signal itself. The spectral shape flexibility and amount of available energy depend on signal length: the longer is the signal, the more flexible and more energy it can deliver. Energy depends on length linearly (different from amplitude, where square law is applicable).

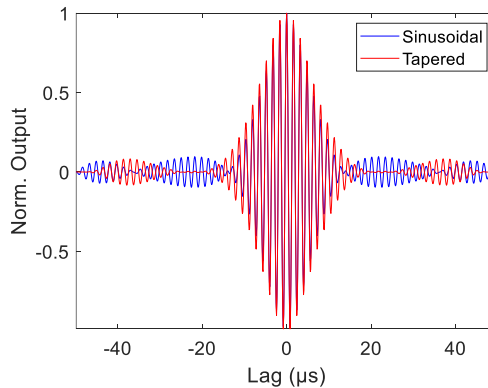


Fig. 12. Autocorrelation of sinusoidal chirp and tapered chirp for sidelobe compression

The important thing to consider here is that results shown are only for excitation signals, thereafter these excitations produce response after passing through the transducer which further reduces the signal bandwidth. Moreover, for arbitrary waveforms a highly linear amplifier and digital to analog converter (DAC) is required. In general, the excitation signal is selected without the accountability of transducer influence, this results not only in the loss of energy, but the correlation and spectral properties remain unpredictable. Moreover, the importance of tapering cannot be denied for the applications of TRUS.

The use of amplitude modulation to compensate for the spectral losses was proposed [92], however, such approach allows to achieve an ideal compensation, but the tradeoff is in the excitation signal energy where full analogue amplifier voltage cannot be exploited. Furthermore, the use of combined amplitude and frequency modulation was also proposed to improve the efficiency of excitation signals [52], however, still there is improvement required for the excitation signal whose spectra could be programmed for desired spectral shape for spectral losses compensation

1.5.1. Bipolar and unipolar excitation signals for transducers

Unipolar excitation signals operate between ground and a positive voltage, contrasting with bipolar signals that swing symmetrically around zero. This fundamental difference yields distinct advantages for unipolar designs, including reduced hardware complexity with a single switch level versus three for bipolar, facilitating integration in portable ultrasound systems. In contrast, unipolar excitation includes a direct current (DC) component that allocates transmitted energy to non-useful low frequencies, potentially halving effective acoustic output and introducing even harmonics that distort the spectrum [89]. Bipolar signals, cancel even harmonics, enhance SNR through full voltage swing, and maintain spectral purity without DC offset, but require more intricate pulsers like push-pull or half-bridge topologies, often necessitating dual supplies and isolated drivers, increasing power consumption and design challenges

The transmitted signal comparison (Fig. 13, left) shows that both excitation types achieve comparable durations and structures, derived from a 100 μ s LFM signal, bipolar exhibits full ± 1 amplitude swing compared to unipolar's 0-1 range, indicating inherent energy superiority in bipolar. The received signal (Fig. 13, right), post system transmission through a modelled air-coupled transducer system (via IIR filter, details in chapter 2), reveals that while the bipolar signal shows slightly higher amplitude and energy density, the unipolar response follows a nearly identical envelope with moderate attenuation. This result highlights that the majority of spectral content is still preserved in the unipolar case, indicating robustness of the excitation strategy even under energy-limited.

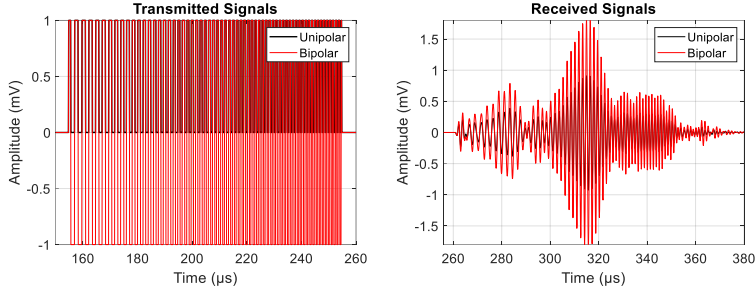


Fig. 13. (left) Transmitted 100 μ s LFM signal and (right) received signal after passing through system model for unipolar and bipolar excitation cases

The spectral SNR comparison between unipolar and bipolar cases is presented in Figure 14. Although bipolar excitation results in marginally higher SNR in the 400–900 kHz range, the overall spectral shape and usable bandwidth remain highly comparable across both excitation types. Unipolar excitation shows flatter behaviour and reduced variance across sub-bands. Specifically, the SNR standard deviation in 400–600 kHz was 11,725 for unipolar and 23,215 for bipolar, indicating nearly twice the fluctuation in the bipolar case. This variance may result from phase distortion, peak signal excursions, or nonlinear interactions exacerbated by higher swing excitation in bipolar transmission.

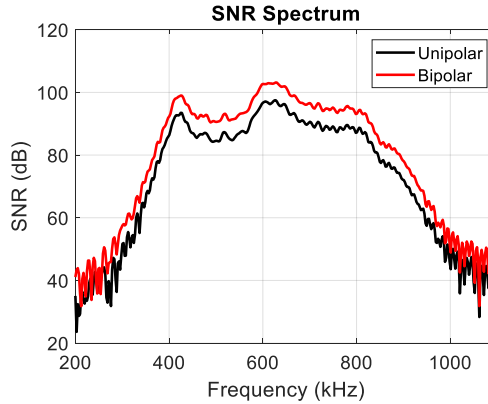


Fig. 14. SNR of received signal using 100 μ s LFM signal after passing through system model for unipolar and bipolar excitation cases

The complexity of the drive circuitry is reflected in the number of switch levels required. Unipolar excitation relies on only one switching threshold ($0 \rightarrow 1$), while bipolar excitation involves three levels ($-1, 0, +1$), requiring dual-polarity driving circuits or high-speed digital-to-analogue conversion. In systems where volume, thermal stability, or simplicity are critical, such as air-coupled ultrasound systems or portable sensing devices, reducing switching complexity leads to tangible hardware advantages. This includes lower heat dissipation, improved timing predictability, and greater integration potential with microcontroller-based platforms. The tabulated results (Table 4) directly support this conclusion.

Table 4. SNR standard deviation with number of switch levels and band of frequency ranges

No of switches: unipolar →1 and bipolar→3		
Bandwidth	SNR Std.	
	Unipolar	Bipolar
200-400 kHz	4454.79	8708.01
400-600 kHz	11725.20	23215.17
600-800 kHz	17027.21	33688.04
800-1000 kHz	7541.22	15032.71

Although bipolar signals may provide a marginal spectral performance edge, the analysis shows that unipolar excitation delivers nearly equivalent received signal quality and significantly more stable SNR across the frequency band. More importantly, it reduces hardware complexity by eliminating the need for DACs or full-bridge amplifiers. In low-power or size-constrained applications, this trade-off is highly favourable.

1.6. Conclusions

The analysis of the literature shows that accurate air-coupled TRUS is fundamentally limited by several interrelated factors: strong frequency and distance dependent attenuation in air, highly non-uniform transducer frequency responses with limited bandwidth, high impedance mismatch and severe SNR degradation at higher frequencies. These constraints currently necessitate the use of multiple transducer pairs and short-pulse excitation, which in turn restricts spectral coverage and measurement accuracy for thin, fragile or moisture-sensitive samples.

Existing spread-spectrum techniques can increase excitation energy without sacrificing bandwidth, yet their long duration causes overlapping reflections that cannot be efficiently separated with conventional methods. Programmable spectral shaping has been proposed to compensate transduction losses in air-coupled TRUS.

Consequently, there is a clear need for a method that (1) compensates spectral losses using optimization, (2) enables high-SNR TRUS with a single transducer pair, and (3) solves the overlapping-reflection problem for long-duration signals. These gaps form the basis for the aim and research tasks of the present dissertation.

2. DESIGN AND OPTIMIZATION OF APWP SIGNALS FOR TRANSDUCTION SPECTRAL LOSSES COMPENSATION

Ultrasonic methods are well recognized for measuring material properties. The use of air-coupled ultrasound is limited in wide range of applications where immersion [93,112] and contact [94] coupling are dominating. However, recent advancements in transduction technology and signal-processing techniques have enabled the use of air-coupled ultrasound for the investigation of thin materials [95–97]. TRUS is used for the simultaneous estimation of velocity and thickness along with attenuation and density by fitting thickness resonance models [98,113]. Nonetheless, the air-coupled ultrasound faces many challenges: acoustic impedance mismatch results in large signal losses [66,99], attenuation in air [100], SNR, and limited bandwidth. Several quarter wavelengths matching layers are used solve the acoustic impedance mismatch [101,102]. Similarly, the modern air-coupled transducers are significantly better in terms of bandwidth and sensitivity improvement [103,104], however, lack of materials for the last matching layer is still limiting the bandwidth, and SNR. The centre frequency of the air-coupled transducer must be kept low due to attenuation in air, as transduction bandwidth is fraction of the centre frequency, thus, attainable bandwidth is not sufficiently broad. Moreover, when the ultrasound wavelength matches the characteristic dimensions of the sample (e.g., thickness $\approx \lambda/2$, λ , $3\lambda/2$), strong resonances occur, resulting in a non-uniform transmission spectrum.

The conventional approach using TRUS is with single spike or pulse signals [96,105,106], due to simple generation electronics and response is easy to interpret and process. The problem with such simple signals is they are highly inefficient, as narrow pulse must be used for broadband excitation, which produces low signal energy, thus, low SNR is obtained. The second problem with pulse or spike signals is that their energy is focused on low frequency (which does not pass the transducer) and therefore less energy is left for high frequencies. There could be improvement done for high SNR by averaging large amounts of signal waveforms [96,105,106], however this is restricted in case of varying parameters (conditions).

The standard approach in TRUS is to use a combination of three or more air-coupled transducers to overcome the limited bandwidth [96,105,106]. However, a single transducer may be used if sufficient SNR is achieved by employing spread-spectrum excitation. SS signals [107] are energy efficient especially rectangular ones [44], in comparison to pulse or toneburst signals, due to their inherent ability of exploiting power source capabilities in full extent. The problem with common SS signals i.e. LFM is that they do not allow control over spectral shape, thus, limits in terms of spectral losses compensation. It was already demonstrated that using amplitude modulation [108], spectral shape of LFM can be altered, however, this does not exploit full power source capabilities and requires DACs, and linear amplifiers, which are not only less efficient in terms of energy but also demand more complex electronics. The solution to such problems can be seen in NLFM [44,107], or APWP signals [49], which offers ability to program the spectral shape for compensation

losses. It was demonstrated [51,53] that excitation spectral shape can be altered to compensate for the transduction losses and increase the available signal bandwidth at the reception's end.

Two issues were identified. First, bipolar rectangular excitation signals were used; these provide more efficient spectral shaping but require more complex excitation electronics compared to unipolar signals. Bipolar signals allow both positive and negative voltage swings, which improves spectral symmetry, increases energy efficiency, and enables better control of pulse compression sidelobes. However, generating bipolar signals requires H-bridge or full-bridge drivers, higher-voltage switching, and careful protection circuitry, increasing the complexity of the hardware. Unipolar excitation, in contrast, is easier to implement because only one polarity is required; the drivers are simpler, cheaper, and more compact, making them suitable for portable or embedded TRUS systems. The trade-off is reduced spectral flexibility and lower effective transmit energy. Second, the noise spectrum was assumed to be flat (white), which is not accurate for piezoceramic transducers. The piezo element of the transducer contributes to TRUS; therefore, its response is more like that of the resonant tank [109]. Moreover, the use of matching and backing layers in transducers introduces extra resonances, thus, impedance of ultrasonic transducers is frequency dependent and noise at the output is not white [110].

To address these issues, a novel approach was proposed based on unipolar excitation signal to attain a transmitted spectrum with uniform SNR for wider bandwidth. Moreover, in comparison to existing studies [51,53], that employ immersion or contact coupling to compensate for spectral losses, both of which inherently provide wide bandwidth, air-coupled ultrasound was used for TRUS. Moreover, as bipolar excitation was used earlier which provides better flexibility in spectral shape programming, the use of unipolar excitation not only simplifies the electronics but also complicates the control over signal spectra. The contributions presented in [36,50] to TRUS are:

- i), single pulse excitation is replaced by spread spectrum signal APWP of programmable spectral shape;
- ii) SNR is improved, there is no need for averaging, which results in more stable results;
- iii) spectral shape optimization, differently from [51,53], involves not just signal losses but also noise spectrum, smooth SNR is obtained over whole bandwidth;
- iv) unipolar APWP sequences are used so simpler excitation electronics can be used.

2.1. TRUS for thin sample in air

TRUS is performed in through-transmission mode using two measurements, calibration (free propagation, Fig. 15, left) and sample (Fig. 15, right). The important thing to consider during measurements is to keep the distance between each transmitting and receiving transducer should be same for all measurements.

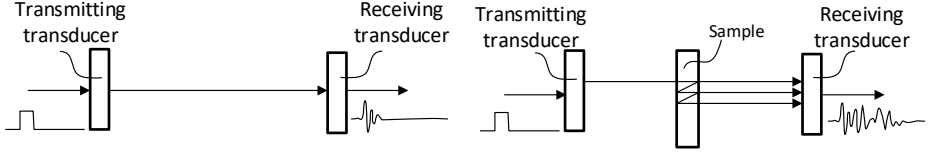


Fig. 15. TRUS in measurement modes: (left) calibration and (right) sample

To investigate the measurements using TRUS, two identical 650 kHz transducers (designed and manufactures by CSIC, Spain) were used in a face-to-face arrangement with a fixed distance of 35 mm between the radiating faces. The sample under test was positioned approximately in the centre of the acoustic path.

The transmission AC (the spectral magnitude and phase response) response $T(\omega)$ (refer Fig. 16 for the example spectrum of the grape (*Vitis vinifera*) leaf) is obtained by taking the ratio between sample and calibration measurements in frequency domain [111,114],

$$T(\omega) = \frac{S_{RXS}(\omega)}{S_{RXC}(\omega)}, \quad (2.1)$$

where, $S_{RXC}(\omega)$ and $S_{RXS}(\omega)$, are the spectrums of calibration and sample signals respectively.

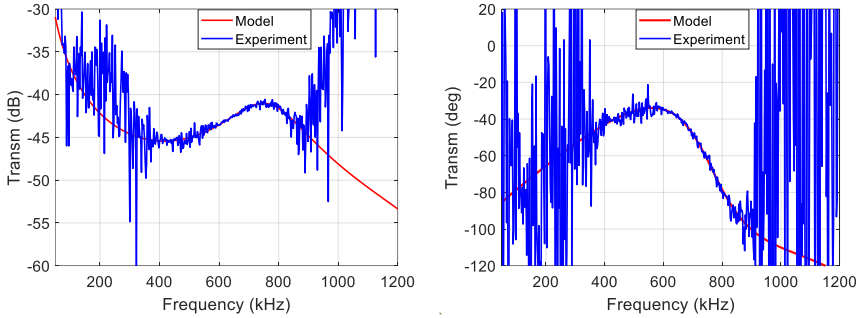


Fig. 16. Transmission response of experimentally measured *Vitis vinifera* leaf, (left) magnitude, and (right) phase response

The resonance frequency can be observed for the *Vitis vinifera* sample at around 780 kHz (Fig. 16, left) due to multiple reflections within the leaf. Through layer modelling of Brekhovskikh's model [98], the propagation of the ultrasonic signal in the sample (i.e. leaf) is described as,

$$T'(\omega) = \frac{-Z_1 \cdot Z_2}{-2 \cdot Z_1 \cdot Z_2 \cos(k' \cdot h) + j(Z_1^2 + Z_2^2) \sin(k' \cdot h)}, \quad (2.2)$$

where, Z_1 is the acoustic impedance of air, Z_2 is acoustic impedance of sample, k' is the complex wave number and h is the thickness of the sample. The complex number is defined as,

$$k' = \frac{\omega}{c_2} - j \cdot \alpha, \quad (2.3)$$

where, α is the attenuation and c_2 is the ultrasound velocity in the sample,

$$\alpha = \alpha_0 \cdot \left(\frac{f}{f_0}\right)^n, \quad (2.4)$$

where f_0 is the normalization frequency which is used as a resonance frequency of transducer.

Once ultrasound velocity and density in sample (c_2, ρ_2) and air (c_1, ρ_1) are known, impedances (Z_1, Z_2) can be calculated as,

$$Z_1 = c_1 \cdot \rho_1, \& Z_2 = c_2 \cdot \rho_2, \quad (2.5)$$

Using the free propagation signal (i.e. calibration signal), sample signal can be estimated as,

$$S_{model}(\omega) = R(\omega) \cdot T'(\omega)e^{j\omega h/c}, \quad (2.6)$$

where, $e^{j\omega h/c}$ is the additional component added in the calculation to account for signal shift due to air replacement with the sample within the thickness h .

$$\mathbf{x} = \arg \min \sum_{i=1}^N \left(\frac{S_{mod}(x)_i - S_{iRXS}}{S_{iRXS}} \right)^2, \quad (2.7)$$

The inverse solution problem (Eq. 2.6) is solved to estimate the sample parameters $\mathbf{x}=(h, c_2, n, \rho_1 \text{ and } \alpha)$ [111]. Through minimization between simulated S_{mod} and experimental S_{RXS} measured signals, inverse solution is solved (Eq. 2.7).

2.2. The need for high SNR and wide bandwidth in TRUS

The transmission spectra (discussed in section 2.2, Fig. 4-5) highlights how each parameter of sample is influencing the response, it is clearly seen that each parameter varied up and down by significant margin individually to reflect how particular parameter affects the spectral shape. This investigation alone provides solid grounds to believe that high SNR (dynamic range) and a broad bandwidth is essential for covering the full range of possible parameter values. If the spectral valley is present at low or high frequencies and due to low SNR, it is hidden in noise, it might implicate in bias errors during the parameter estimation. The important thing to consider here is that even if the spectral peak (resonance peak) falls out of the passband region, bias errors can be kept low if there is sufficient SNR for the rest of the spectra. The transmission response (Fig. 16) was obtained using pulse excitation signal (blue curve), which highlights that high SNR is only possible around the centre frequency of transducer in very narrow bandwidth.

The resonant nature of piezo element and matching layers used for bandwidth and sensitivity improvement produces the non-uniform spectral response (Fig. 17), although using spread spectrum signals, i.e. LFM (Fig. 17, black), signal level can be increased in comparison to the pulse (Fig. 17, red), but still the bandwidth covered is insufficient.

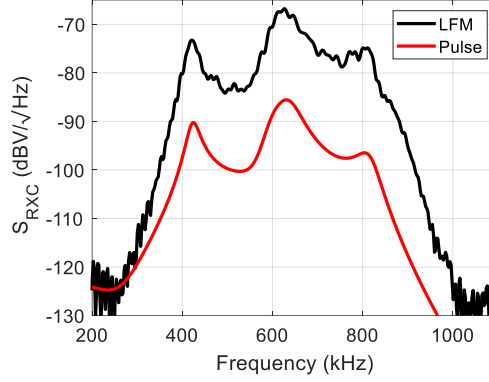


Fig. 17. The magnitude spectrum of the received signal when pulse (red) is used as excitation signal and LFM (black) signal for air-coupled transducer

The signal spectrum can be flattened (for better efficiency) or bandwidth could be increased by using APWP signals [49] with ability of programmable spectral shape (Fig. 18).

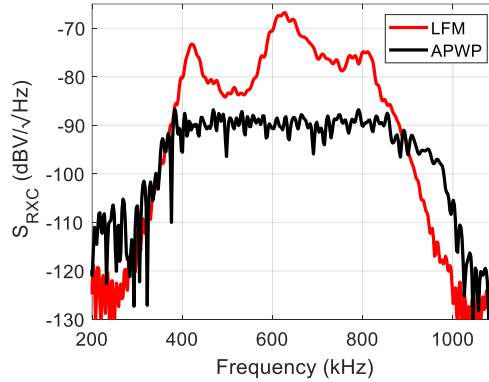


Fig. 18. The magnitude spectrum of the received signal when air-coupled transducer is excited by LFM (red) and losses-compensating APWP signal (black) signal

The compensation could be made to increase bandwidth by pushing the energy from within the passband (i.e. around centre frequency where energy is already high) towards the passband edges at the expense of some signal energy loss. This is evident in APWP case (Fig. 18), where improvement can be observed at 900-1000 kHz. However, as the transducer impedance is complex in nature [110], therefore commonly it is assumed that noise at the output amplifier is flat i.e. white. Therefore, only signal spectrum is compensated, which is not accurate approach as in piezoceramic, especially in air-coupled transducers. The noise at the receiving transducer output can be calculated by putting highly attenuated sample between the transducers which will limit any signal to pass through, thus, only thing received at output transducer is therefore noise, as shown in Fig. 19. The output noise and

preamplifier gain AC response was measured first, afterwards results were converted to input noise voltage spectral density (NVSD) [110].

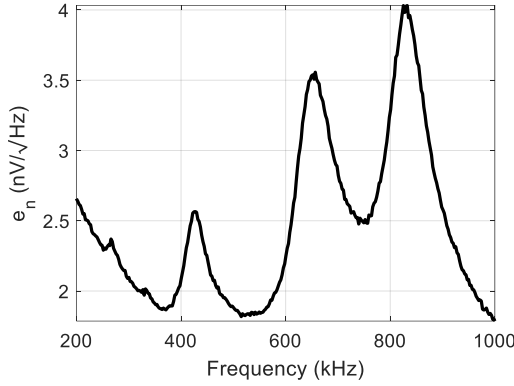


Fig. 19. Preamplifier input NVSD when connected to receiving transducer

It can be noticed that noise spectral density varies across the passband of the transducer: at 400 kHz it is 2.65 nV/√Hz and at 830 kHz it is 3.45 nV/√Hz. Therefore, despite receiving signal spectrum being flattened, the resulting SNR is not flat (see Fig. 20).

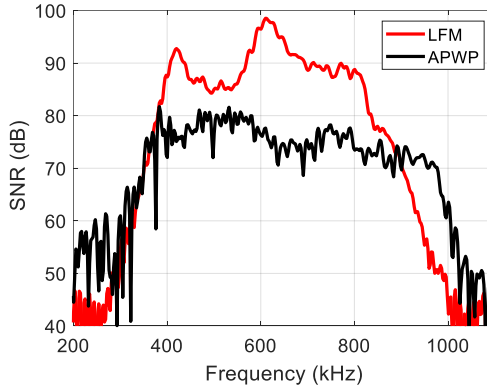


Fig. 20. Received signal SNR when transducer is excited by LFM (red) and losses compensating APWP (black) signal

Thus, it is observed, that in comparison to immersion or contact ultrasound, air-coupled transducers are more complex mainly due to complicated construction, and narrower bandwidth. Therefore, it is examined that flattening the transmission response does not provide flat SNR for TRUS measurements.

The waveform is mainly dependent on the form of excitation signal, it is established that APWP signals offer best results, but the main problem is DAC and linear analogue amplifiers. This not only increases the cost of equipment but also results in inefficient equipment. Rectangular excitation signals on the other hand require simple switches, thus removing the need for DAC or linear analogue

amplifiers. The use of bipolar excitation was already done in existing studies [51,53], which provide more flexibility for spectral shape programming, however complex equipment is also needed. A unipolar excitation signal was proposed [36,50], which simplifies the equipment requirement to produce APWP signals and efficiency, and cost is improved furthermore. Thus, duration and position of each pulse can be altered for desired APWP response (Fig. 21).

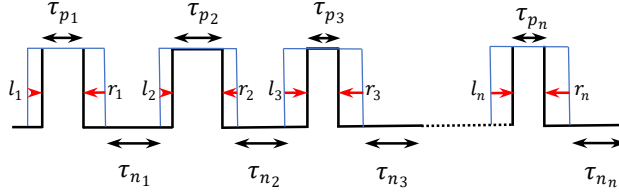


Fig. 21. A unipolar excitation signal for APWP

The problem with such is unipolar signals produce a spectral ripple due to convolution with *sinc* function; moreover, it has less control over signal spectra, therefore, the optimization process becomes more complicated.

2.3. Unipolar excitation for flat SNR

The existing studies [51,53] focused on the development of APWP based optimization approach, however, it is observed that it is SNR that should be used for optimization instead of only signal. Moreover, if the signal spectrum is flat, SNR spectrum is not flat due to frequency dependent noise. The better approach would have been to compensate for the sample signal; however, this would have required new APWP signal optimization for every measurement. Therefore, spectral losses compensation was developed for calibration setup. This would allow one time SNR based APWP optimization only for all measurements. The corresponding compensation path is shown on Fig. 22.

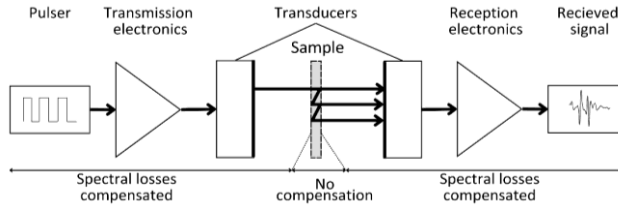


Fig. 22. SNR compensation for spectral losses signal path

The spectral losses compensation was done only for the components that were influencing the measurements, i.e. electronics, transducers, and pulser, the benefit of using SS signals is that they can be made long for better flexibility and therefore improves the SNR without implicating the bandwidth. Therefore, the high SNR (around centre frequency of transducer) can be sacrificed to achieve spectral losses compensation that will flatten the SNR AC response.

NLFM signals provide the ability of spectral shape programming; thus the derivation of APWP signals starts from NLFM signal production. Optimization is achieved for free propagation (calibration) signal by ensuring SNR flatness with selected frequency range.

The NLFM signal was derived by first obtaining the system transmission AC response using broadband (300-950 kHz) 100 μ s chirp (LFM). For better statistics to compensate for environmental and electronics noises, multiple waveforms ($M = 1000$) of signal s_{RXC} were obtained at the output of reception amplifier. These were thereafter averaged in time domain and resultant averaged signals were transformed to frequency domain for further processing,

$$S_{RXC} = \sqrt{\frac{2 \cdot DFT(\sum_{i=1}^N s_{iRXC})^2}{N \cdot f_s \cdot M}}. \quad (2.8)$$

As it was proposed to develop spectral losses compensation for SNR based optimization, therefore noise spectral density was measured next. The noise spectral density was measured by placing high attenuating material as sample between the transducer so that no signal is transmitted, therefore at the receiving transducer output only thing received is noise. Like transmission response, M measurements were obtained, transformed into frequency domain and averaged power-wise to obtain NVSD of sampling frequency f_s and number of samples N :

$$e_n = \sqrt{\frac{2 \cdot \sum_{i=1}^N |DFT(s_{iRX0})|^2}{N \cdot f_s \cdot M}}, \quad (2.9)$$

The main objective is to obtain SNR for the derivation of compensation path:

$$SNR_C = \frac{S_{RXC}}{e_n}. \quad (2.10)$$

The compensation function is obtained as follows:

$$\kappa(f) = \frac{|W(f)|}{|SNR_C(f)|}, \quad (2.11)$$

where, $W(f)$ is the spectral windowing function used to control the bandwidth. The windowing function was achieved using *Tukey* window with decay of 0.3 from 300-950 kHz.

The compensation function is developed for excitation signal (Fig. 23), particularly the LFM signal, the desired excitation signal for flat spectrum is:

$$S_{TXdes}(f) = \kappa(f) |S_{TX}(f)|, \quad (2.12)$$

where, $S_{TX}(f)$ is the spectrum of excitation signal. Similarly, for NLFM such spectrum could be obtained where energy of each spectral component is determined by the rate of change of instantaneous frequency.

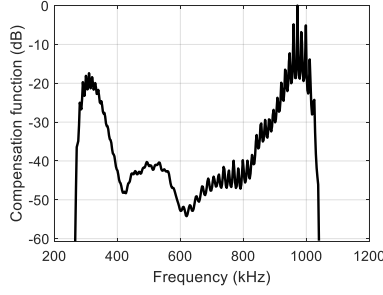


Fig. 23. Compensation function

The derivation of NLFM is possible by first achieving time-frequency relationship (Fig. 24), as:

$$\tau(f') = \int_0^{f'} (\kappa(f)|^2) df. \quad (2.13)$$

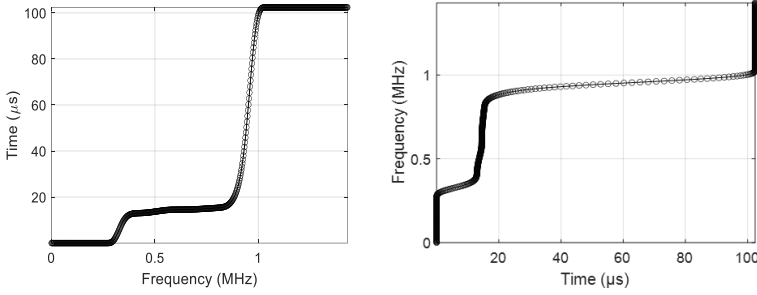


Fig. 24. (left) Time-frequency characteristic function of compensated NLFM signal, (right) interpolated instantaneous frequency function

For the instantaneous phase calculation, expression in Eq. 2.13 needs to be converted to $f(\tau)$ (Fig. 24, left), therefore by resampling non-uniform grid τ :

$$\tau(f') \xrightarrow{\text{Interp}} f'(\tau). \quad (2.14)$$

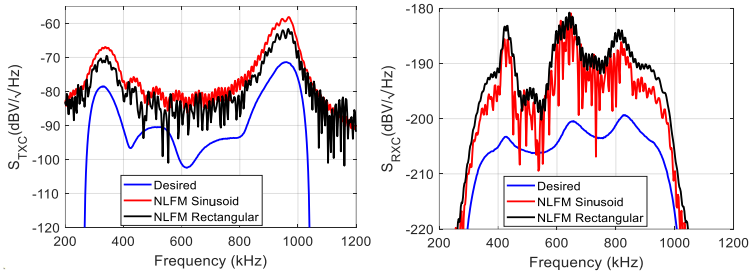


Fig. 25. (left) Comparison of transmitted and (right) received signals spectra for the desired (blue), sinusoid NLFM (black) and rectangular NLFM (red) signals after spectral losses compensation. Note that the desired spectral shape was not obtained

Thus, phase is obtained by integration over the instantaneous frequencies:

$$\theta(t') = 2\pi \int_0^{t'} f'(t)dt, \quad (2.15)$$

The required NLFM signal is thereafter obtained:

$$s_{TXNLFM}(t) = \text{sign}(\sin \theta(t)). \quad (2.16)$$

The advantage of using NLFM signal is that it effectively utilizes the output range of amplifiers as envelopes are constant over time. Unipolar excitation is converted from bipolar (sign operation) to produce rectangular waveform, which results in much simpler electronics requirement. Received signal S_{RXC} spectra for the desired (+/-1 V) and unipolar NLFM (+1 V) signals are shown in Fig. 25, where it is clear the spectral compensation increases the bandwidth but still the spectral shape is not as desired.

To achieve the desired shape, conventionally proper envelope modulation is applied and revert to sinusoidal signal, this means turning to amplitude modulated-NLFM signals. Which would require arbitrary waveforms generator in the form of linear amplifiers, and generators, thus system would become bulky and costly. For better voltage swing dynamic utilization and improved efficiency, unipolar excitation signals were used. The APWP sets are proposed for final spectrum shaping due to their ability of spectral adjustment using pulse duration and placement in time.

The problem with APWP is lengthy optimization process. To speed up the optimization process, a numerical model of the system transmission response of calibration measurement was developed using 8th order infinite impulse response (IIR) filter from the experimental signal as close simulated approximation. The distance between fixed receiving and transmitting transducers was 35 mm, excitement using 700 ns duration and 50 V pulse. Average of 1000 waveforms were obtained for sufficient SNR of calibration signal S_{RXCexp} . The subsample delay $e^{-j\omega ToF}$ equal to ToF was added in IIR estimation:

$$T_{sys} = \frac{S_{RXCexp}}{(V_{HV} \cdot P \cdot e^{-j\omega ToF})}, \quad (2.17)$$

where P is the excitation signal in the form of 0's and 1's, V_{HV} is the amplitude of excitation signal, T_{sys} is the system (including electronics, transducers) transmission AC response. The delay operator was added to compensate for propagation in air, and partial delay in transducers and electronics. The system transmission AC response was then inverted to get IIR coefficient's a', b' (using MATLAB *invfreqz* function). These coefficients were used as upper and lower bound constraints for particle swarm optimization (PSO) derivation:

$$LB(a, b) = (a', b')/2 \text{ \& } UB(a, b) = (a', b') * 2. \quad (2.18)$$

The convergence criteria was reminder energy between simulated and experimental measured signals for the optimization of IIR coefficients:

$$y = \arg \min \sum_{i=1}^N \left(\frac{S_{RXCsim_i} - S_{RXCexp}}{S_{RXCexp}} \right)^2, \quad (2.19)$$

where s_{RXCsim} is the simulated calibration signal produced as

$$s_{RXCsim} = IIR(P, a, b) \cdot e^{-j\omega \cdot ToF}, \quad (2.20)$$

ToF was estimated as a cross-correlation peak between $IIR(P, a, b)$ and P with cosine subsample interpolation. Optimized parameters and ToF were saved as the calibration propagation model coefficients. To further improve the stability, second order section (SOS) was used after IIR coefficients.

The conversion of SNR-compensated NLFM signal into APWP signal was started from direct assignment (APWP = NLFM). Simulated signals (Eq. 2.20) were used to speed up the lengthy optimization process. The $SNR_{Ci}(f)$ was obtained from excitation signal P_i , generated in the i^{th} iteration. SNR spectral flatness ξ was evaluated within the desired passband $f_{min} \sim f_{max}$:

$$\xi_i = \sqrt{\frac{\int_{f_{min}}^{f_{max}} \|\overline{SNR_{Ci}(f)} - |\overline{SNR_{Ci}(f)}|\|^2 df}{|\overline{SNR_{Ci}(f)}|}}. \quad (2.21)$$

The SNR flatness was regulated by adjusting the individual APWP pulses duration and position in time.

The optimization process is described as:

1. Transmit the original APWP signal and estimate the initial spectral flatness $\hat{i}_0$. The APWP signal contains N rectangular pulses, each L_n samples long ($n=1, \dots, N$);
2. Resize the n^{th} pulse (starting from the first), transmit it and estimate the new spectral flatness ξ_{in} . Resizing (Fig. 21) involves two secondary loops, outer one which adjusts the pulse's left side by $l = -L_n/2, \dots, 3L_n/2$ and inner one adjusts the right side by $r = -3L_n/2, \dots, L_n/2$;
 - a. Resize the n^{th} pulse left size by l ;
 - i. Resize the n^{th} pulse right side by r ;
 - ii. Transmit the resized pulse and use the received signal to calculate the SNR spectral flatness $\hat{i}_{lr}$;
 - b. Select the best contender from $l \times r$ ξ matrix (the one that provides minimum ξ) and set $\xi_i = \min_{l,r} \xi_{lr}$;
3. Replace the n^{th} pulse with the one with best ξ and repeat 2 for the next pulse updating $n=n+1$ until $n=N$;
4. Calculate the spectral flatness reduction of the i^{th} iteration $D\xi_i = \xi_i - \xi_{(i-1)}$. Repeat 2 until $D\xi_i$ lays below certain threshold or until maximum iterations are reached.

Envelope reduction results in pulse width reduction of the APWP signal. The interpolation errors introduced from Eq. 2.14 were reduced due to pulse position shifting.

Time domain excitation signals in case of either NLFM or APWP does not reflect visible information (Fig. 26, left and centre), thus spectral domain is preferred. However for comparative analysis, the APWP signal that was derived from NLFM

signals is shown in Fig. 26 (right) by comparing the part of full signal showing the difference between each pulse before and after compensation.

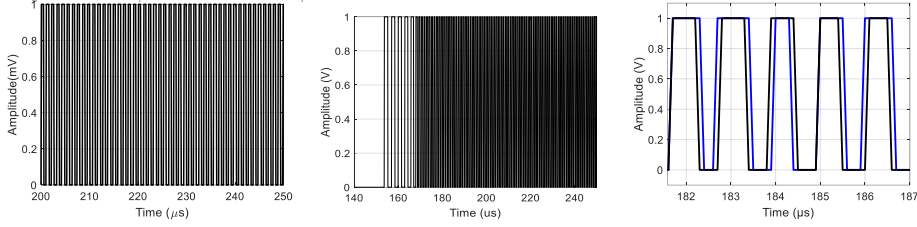


Fig. 26. (left) Time domain unipolar NLFM signal of 100 μ s, (centre) time domain unipolar APWP signal of 100 μ s, (right) time domain segment of APWP pulse train for unipolar case: black not compensated, blue compensated

Using proposed scheme APWP is used for the corrective measurement and full permutation of each pulse position and duration values. Global convergence is not fully satisfied with full permutation; therefore, several iterations are required to ensure that global optimum is achieved. The optimization process performance can be seen from SNR spectral flatness improvement for each iteration (Fig. 27, left) and how much improvement is achieved in every iteration (Fig. 27, right).

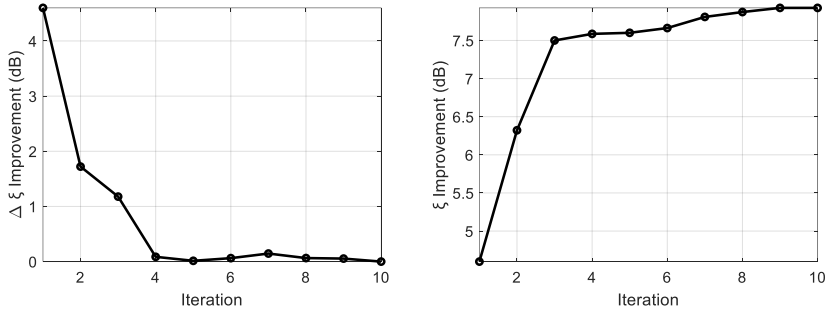


Fig. 27. (left) SNR spectral flatness improvement ($\Delta\xi$) for each iteration and (right) total improvement (ξ) in each iteration

The result can be evaluated by comparing the SNR spectral flatness in Fig. 28. APWP produces the best SNR spectral flatness compared to pulse, LFM, and NLFM. The results displayed were all generated using the same peak amplitude for all transmitted signals.

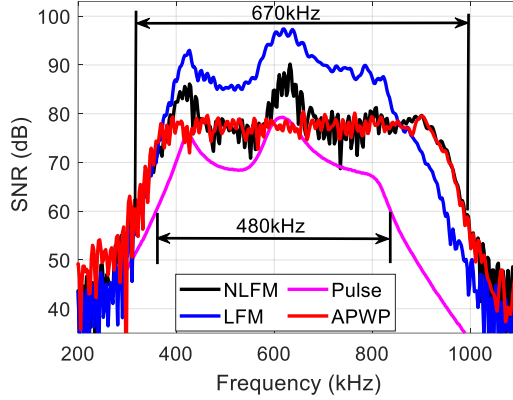


Fig. 28. The received SNR comparison. Note that SNR is flat only for APWP signal. Bandwidth was improved for NLFM and APWP signals

It can be noted that SNR-equalized APWP provides improved spectral losses compensation at frequencies beyond 900 kHz and below 380 kHz. The -20 dB bandwidth for pulse was 480 kHz, meanwhile APWP was able to achieve 670 kHz bandwidth. On the other hand, NLFM provides similar bandwidth, but the SNR spectrum is not as flat. Additionally, APWP provides improved SNR.

2.4. SNR flatness evaluation on TRUS

Experimental and simulated evaluation was achieved for the proposed approach; simulation was developed for grape (*Vitis vinifera*) leaf (using the approach described in section 2.3, i.e. Eq. 2.20) and experimental evaluation was achieved using thin 2.045 mm polycarbonate plate. The focus was to evaluate bias errors for the estimation of parameters. Through mechanical measurements, only parameters obtained can be thickness and density, in case of plant leaves it is difficult to estimate the thickness due to fragile nature of leaf, also living organism exhibits varying properties over time depending on the environmental conditions. Complex anatomy of leaf makes it also difficult to estimate density, circadian rhythm changes the properties of leaf depending on water level, temperature, and lighting. Thus, only numerical simulation was achieved for *Vitis vinifera*.

The experimentally obtained values for plant leaf are $a = 800 \text{ Np/m}$, $c = 315 \text{ m/s}$, $h = 250 \text{ }\mu\text{m}$, $\rho = 890.89 \text{ kg/m}^3$ from Eq. 2.2-2.5. Simulation results in noise free signal, therefore, random noise was added, for each of 1000 signal waveforms (statistically following the Fig. 19 spectral shape) Gaussian noise.

The inverse solution for TRUS (Eq. 2.1-2.5) was achieved using PSO with objective function described in Eq. 2.7. The estimated sample parameters were averaged and then compared with actual (best available estimate) measurements for bias error calculation. Using constrained boundaries fitting process was achieved with lower boundary as half of the average estimated parameters and upper boundary as

double of the average estimated parameters (Eq. 2.18). The transmission response of each of Pulse, LFM, NLFM, and APWP are presented in Fig. 29.

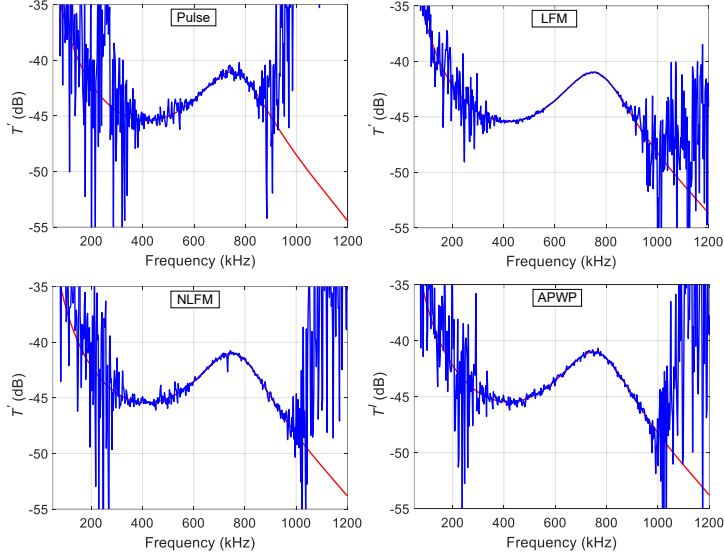


Fig. 29. Transmission magnitude for *Vitis Vinifera* leaf in case of different excitation signals: (top left) pulse, (top right) LFM, (bottom left) NLFM and (bottom right) APWP

The performance of each excitation signal can be clearly seen through transmission magnitude spectrum. Pulse being the simplest form of excitation offers smallest energy; thus, SNR is worst (Fig. 29, top left) as bandwidth coverage is only 400-820 kHz. Highest energy can be seen for LFM with best SNR and band coverage from 380-880 kHz (Fig. 29, top right). However, it was assumed that SNR flatness is more important than large energy at frequencies where transduction losses are small. Therefore, energy was pushed into stopband for NLFM and APWP signals. NLFM and APWP provided acceptable SNR at broader bandwidth of 300-950 kHz (Fig. 29, bottom left and bottom right respectively), compared to LFM. Yet, small distortions are visible in the passband of NLFM and APWP spectra as energy was removed from this area.

2.4.1. Experimental setup: excitation and reception electronics

The experimental setup used for the evaluation of polycarbonate plate is shown in Fig. 30, using air-coupled 650 kHz centre frequency transducers mounted at fixed distance of 35 mm. The 2.045 mm polycarbonate plate was mounted using linear actuator for automated insertion and removal mechanism of sample measurement. The whole experiment was conducted in thermal isolation box made from polystyrene to consequently reduce the ambient temperature variation. Registered temperature range inside the box during whole experiment time ranged from only 24.13 °C to 24.30 °C (an approximate variation of only 0.704%).

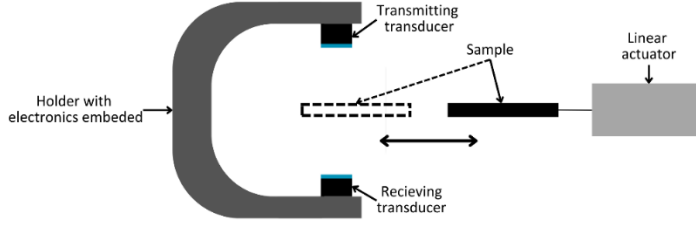


Fig. 30. Experimental setup used for the evaluation of polycarbonate sample by using air-coupled ultrasonic data acquisition system

The hardware electronics was adopted from already established work [29,35,54], preamplifier with capacitive stabilization (Fig. 31) was used to ensure accurate signal detection in air-coupled measurements, which is critical at the receiver stage. The design was optimized for low input-referred noise and bandwidth suited for ultrasonic frequencies. The circuit was configured in an inverting topology with feedback stabilization. Capacitors C_1 – C_6 serve to filter DC offsets and stabilize power supply rails, while resistors R_1 and R_2 define gain and bandwidth. This topology is designed to maintain linearity and minimal distortion while preserving signal fidelity. Importantly, understanding the noise contributions of each passive and active element in this configuration is vital for quantifying system limitations, especially in low-amplitude ultrasonic multiple reflection cases [35].

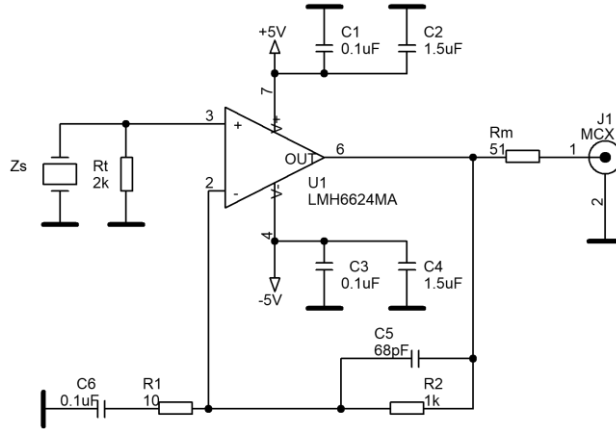


Fig. 31. Low-noise inverting preamplifier using LMH6624MA with capacitive stabilization [35]

The system-level design and the internal electronics governing signal generation and acquisition is presented in Fig. 32 where a dedicated high-voltage (HV) DC-DC converter was used to supply the transmitter. The signal travels through the transducer pair and received, then amplified, digitized, and processed by an embedded microcontroller unit (μ C). Power management system (PMS) and Bluetooth low energy (BLE) modules ensure standalone operation and wireless communication.

Notably, synchronization and buffering are managed by acquisition control (ACQ sync & buffer), ensuring precise timing required for phase-sensitive analysis

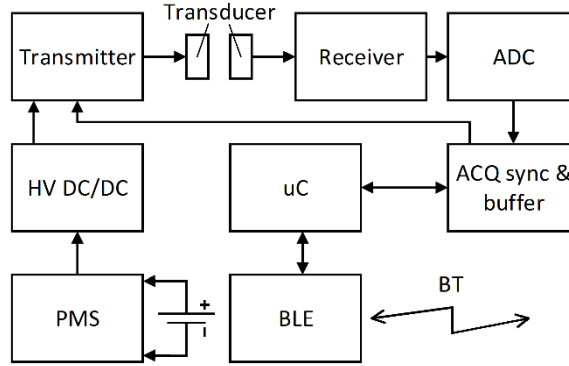


Fig. 32. System-level block diagram of a portable air-coupled ultrasonic measurement system [29]

At the core of the excitation circuitry lies the pulser stage, responsible for generating high-voltage (unipolar or bipolar signals). Fig. 33 depicts a push-pull high-voltage pulser circuit comprising two high-speed MOSFETs controlled through isolated gate drivers. The use of bootstrap capacitors allows for efficient voltage swing from 0 to $\pm HV$, enabling sharp edge transitions essential for short-duration pulses like those required for coded excitation methods. Although originally designed for bipolar signalling, this topology can be adapted for unipolar operation by fixing one of the legs or modifying the logic level inputs to restrict the output swing between 0 and $+HV$.

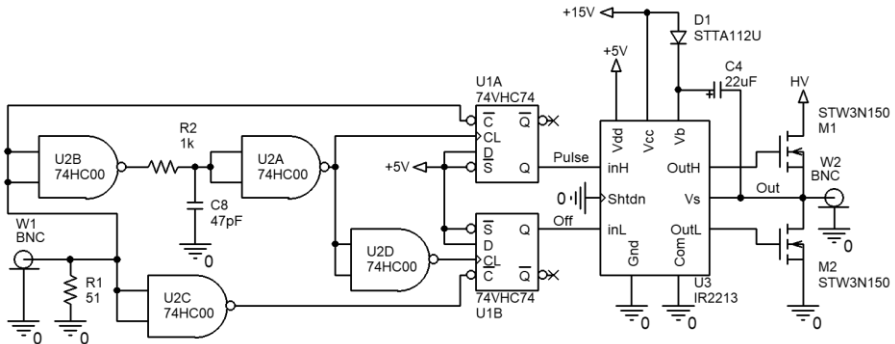


Fig. 33. Bipolar pulser using high and low side driver [54]

The magnitude transmission response (using Eq. 2.1) is shown in Fig. 34, the experimental transmission response (blue) is compared to the inverse solution (red) result for each of the four excitation signals (pulse, LFM, NLFM and APWP).

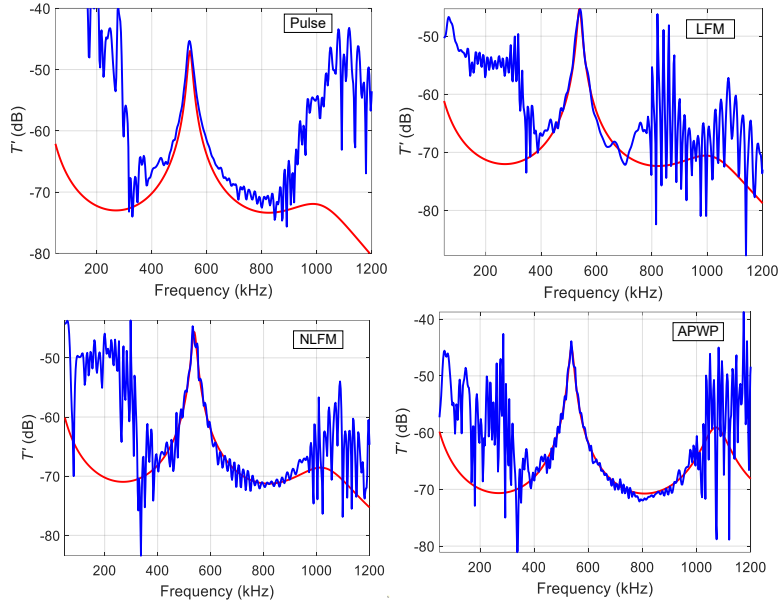


Fig. 34. Transmission magnitude for polycarbonate plate in case of different excitation signals: (top left) pulse, (top right) LFM, (bottom left) NLFM and (bottom right) APWP

The actual thickness of the sample is 2.045 mm and estimated ultrasound velocity is 2240 m/s, this corresponds to resonance peak at 537 kHz. It can be noted that inverse solution results for pulse and LFM signals differ from experimentally obtained transmission so large bias errors can be expected for these signals. Acceptable SNR in the case of pulse excitation is ensured only for 450-620 kHz. For LFM excitation 400-700 kHz range is covered reliably, 380-930 kHz range is covered by NLFM. The best coverage is by APWP, 380-1000 kHz. Final judgement should be based on sample parameters (sample thickness, density, ultrasound velocity and attenuation) estimation performance.

2.5. Sample parameters estimation performance

The sample parameters were estimated using inverse solution for TRUS (explained in section 2.1 and section 2.4 using Eq. 2.2-2.5), as the waveform for experiment and simulation were 1000; therefore, resulting solution provides 1000 values for sample thickness, ultrasound velocity, density and attenuation. The measured values were considered as actual value which was average of all 1000 values for each parameter, in case of *Vitis vinifera* the reference values used were: $a = 800$ Np/m, $c = 315$ m/s, $h = 250$ μ m, $\rho = 890.89$ kg/m³ (obtained by averaging measurements from 10 leaves, with 1000 repeated measurements acquired for each leaf). For polycarbonate samples, reference values were calculated using caliper and weight measurement ($a = 43.5$ Np/m, $c = 2240$ m/s, $h = 2.045$ mm, $\rho = 1193.6$ kg/m³), along with literature standard for polycarbonate values for correct

estimation [43]. Relative bias errors were obtained by taking the ratio of absolute error to reference value.

Bias error estimated a for *Vitis vinifera* (Fig. 35, left) and polycarbonate (Fig. 35, right) samples provide comparison where it can be noted that largest bias errors are for LFM excitation (16%), followed by pulse (12%), NLFM (5.5%) and using SNR optimized APWP errors are least (1%) for *Vitis vinifera*. Similarly, the experimental results using polycarbonate also outperform in case of APWP (17%), slightly higher than simulated, however it was expected due to environmental influences.

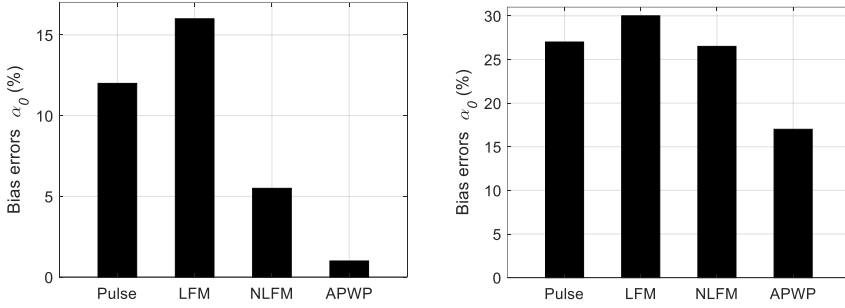


Fig. 35. Relative bias error attenuation a for (left) *Vitis vinifera* and (right) polycarbonate

Relative bias errors for c are shown in Fig. 36, for *Vitis vinifera* (Fig. 36, left), highest errors are reported for NLFM (7%), followed by pulse (4.5%) and LFM (2.4%), APWP results in lowest error of only 0.3%. The bias errors for polycarbonate case (Fig. 36, right), are highest for pulse (2.9%), and least for APWP (0.2%).

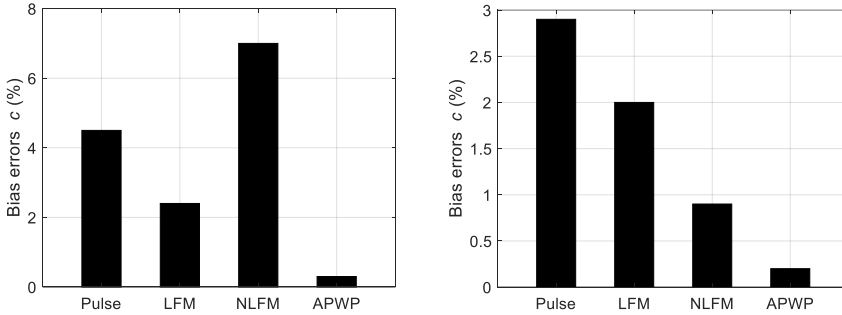


Fig. 36. Relative bias error velocity c for (left) *Vitis vinifera* and (right) polycarbonate

The bias errors for r (Fig. 37) and h (Fig. 38) are also least for the APWP in both simulated and experimental cases. Therefore, through experimental evaluation it can be concluded that spectral losses compensation with SNR equalization reduces the bias errors in TRUS measurements.

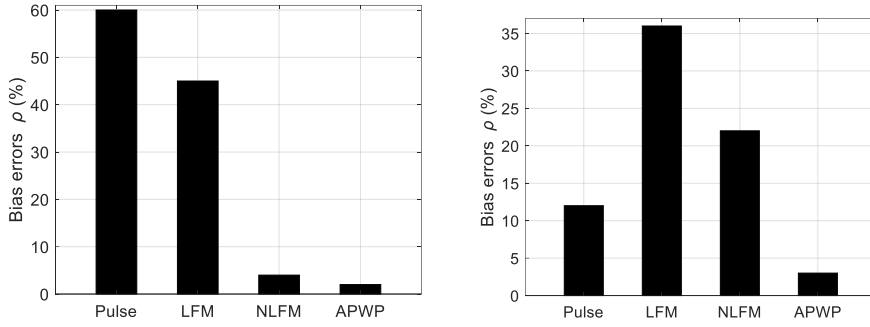


Fig. 37. Relative bias error density ρ for (left) *Vitis vinifera* and (right) polycarbonate

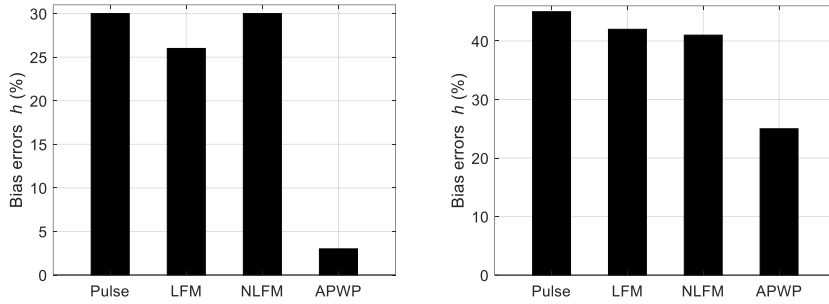


Fig. 38. Relative bias error thickness h for (left) *Vitis vinifera* and (right) polycarbonate

The relative bias errors for the four acoustic parameters (a , c , ρ , and h) show material-dependent behaviour between *Vitis vinifera* leaves and polycarbonate (Figs. 35–38). For *Vitis vinifera* (left panels), the highest bias were observed for ρ (up to 60 %) and h (up to 30 %), while for polycarbonate (right panels) the largest errors appear in a (up to 30 %) and c (up to 3 %).

This pattern is physically consistent with the acoustic properties of the two media. Polycarbonate (2.05 mm thick, high impedance, low attenuation) produces strong multiple internal reverberations that overlap with the primary reflections, increasing spectral ripple and corrupting velocity and attenuation estimates. In contrast, grape leaves are thin, highly attenuative, and biologically inhomogeneous, leading to weaker secondary reflections but greater variability in reflection coefficient and thickness estimation due to surface roughness and local density fluctuations.

The model therefore exhibits graceful degradation: parameters that depend strongly on reverberation suppression (a and c) are more affected in polycarbonate, whereas parameters sensitive to local interface properties (r and h) show higher bias in the leaf samples. These results highlight that the residual error is not random but directly traceable to the dominant physical mechanisms in each material.

2.6. Conclusions

A SNR-optimized APWP signal with a programmable spectral form was proposed to guarantee high SNR within the transducer passband. Compared with conventional pulse, APWP signals enable significantly wider bandwidth (670 kHz vs. 480 kHz at -20 dB) without increasing peak voltage, while unipolar excitation further simplifies the driving electronics..

The performance of the proposed approach was evaluated through simulation on *Vitis vinifera* leaves and experimental measurements on a 2.05 mm polycarbonate plate. In both cases, the SNR-optimized APWP signal produced the lowest bias errors for all four estimated parameters (a , c , ρ , and h). For the leaf simulation, APWP yielded the smallest relative bias errors across all parameters, while LFM consistently showed the largest deviations. In the polycarbonate experiment, APWP again outperformed the other excitations, with the largest errors observed for pulse and LFM signals.

These results demonstrate that spectral compensation using APWP not only increases usable bandwidth but also provides the most accurate and robust parameter estimation in air-coupled ultrasonic testing.

3. STUDY OF SPECTRAL CONTENT EFFECTS IN NON-CONTACT RESONANCE ULTRASOUND SPECTROSCOPY

APWP signals were proposed for transduction losses compensation and performance analysed when SNR equalization is used in previous section. In previous analysis (Chapter 2), energy is pushed away from spectral components where SNR is already good towards the transition bands of the transducer to increase the bandwidth. However, the optimization requirements for the TRUS still remain unclear. This section focuses on the development and comparison of several optimization criteria for spectral shape programming. Another aim is to analyse what spectral coverage should be relative to the resonance peaks and valleys of the sample for lowest measurement errors.

Six different APWP optimization convergence criteria were explored, i) aimed at SNR flatness, ii) aimed at LFM SNR improvement on a linear scale, iii) aimed at LFM SNR improvement on a linear scale with the SNR as an additional weight, iv) aimed at LFM energy improvement on a dB scale, v) aimed at NLFM SNR improvement on a dB scale and vi) aimed at NLFM 1/SNR improvement on a dB scale. Each criterion was focused only on desired frequencies, $f_{min}f_{max}$.

The compensation function was developed using SNR_c , the SNR AC response of the current version of APWP signal, calculated using (3.1).

Criteria (i), aimed at SNR flatness, is based on standard deviation of SNR_c the SNR AC response of the current version of APWP signal within the desired passband $f_{min}f_{max}$. That is APWP signal is optimized in such way that energy is pushed towards the edges of the passband where losses are high.

Criteria (ii), aimed at SNR improvement on a linear scale. LFM signal (instead of NLFM) was used as initial guess. That is, APWP signal is optimized in such way that energy is attempted to improve over LFM signal, no matter which frequency.

Criteria (iii) aimed at LFM SNR improvement on a linear scale as per (3.3), but with the SNR as an additional weight. LFM signal (instead of NLFM) was used as initial guess. That is, APWP signal is optimized in such way that energy is attempted to improve over LFM only for frequencies where spectral losses are high.

Criteria (iv) aimed at LFM energy improvement on a dB scale. LFM signal (instead of NLFM) was used as initial guess. That is, APWP signal is optimized in such way that energy is attempted to improve over LFM, no matter which frequency, same as (3.3) only with dB sensitivity.

Criteria (v) aimed at NLFM SNR improvement on a linear scale. NLFM signal was used as initial guess. That is, APWP signal is optimized in such way that energy is attempted to improve over NLFM signal no matter which frequency.

Criteria (vi) aimed at NLFM 1/SNR improvement on a dB scale. NLFM signal was used as initial guess. That is, APWP signal is optimized in such way that energy is attempted to degrade over NLFM, no matter which frequency.

3.1. Air-coupled resonance ultrasound spectroscopy

TRUS for air-coupled ultrasound requires two measurements, calibration and sample in through transmission mode (Fig. 39). The transmission response is obtained in frequency domain by taking the ratio of sample and calibration measurements. The difference between the sample and calibration signals is the propagation within the sample effects. Then, the sample parameters can be extracted from the transmission response which can be obtained by calculating the ratio between the sample and calibration spectrum.

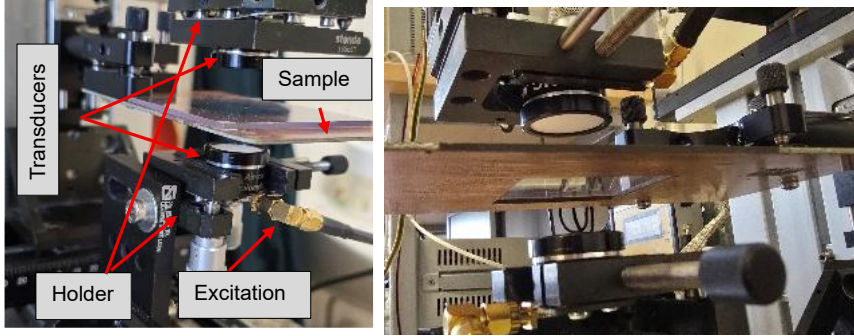


Fig. 39. Air-coupled measurement setup

In comparison to the frequency domain fitting of results [115], the simpler and efficient time domain fitting is proposed which requires gating of the reflections as convolution of the *sinc* function results in the ringing in transmission spectra if the gate cuts out segment of the received signal, thus, fitting quality is reduced significantly.

The sample signal S_{mod} is derived from the measured (experimental) calibration signal R , using transmission response model T :

$$S_{mod}(\omega, \mathbf{y}, \mathbf{x}) = T(\omega, \mathbf{y}, \mathbf{x}) \cdot S_{cal}(\omega) \cdot e^{j \cdot \omega \frac{h}{c_{air}}}, \quad (3.1)$$

where, $T(\omega, \mathbf{y}, \mathbf{x})$ is the transfer function that depends on the air parameters ($\mathbf{x}$) and sample parameters ($\mathbf{y}$). $S_{cal}(\omega)$ is the calibration signal with $\omega = 2\pi f$ as the angular frequency, h is the thickness of the sample and c_{air} is the ultrasound velocity in the air.

The transmission response used is demonstrated by Berkhovskikh's study [98], for ultrasound through sample Eq. 2.2-2.7 (Chapter 2 already describes these equations).

The modelling results in comparison to the measured signals are demonstrated in Fig. 40 (left). The results are good in terms of fitting and even remainder (Fig. 40 right) is quite low indicating a high SNR (good sample resonance peak match to transducer passband, low sample attenuation and sample impedance close to that of air).

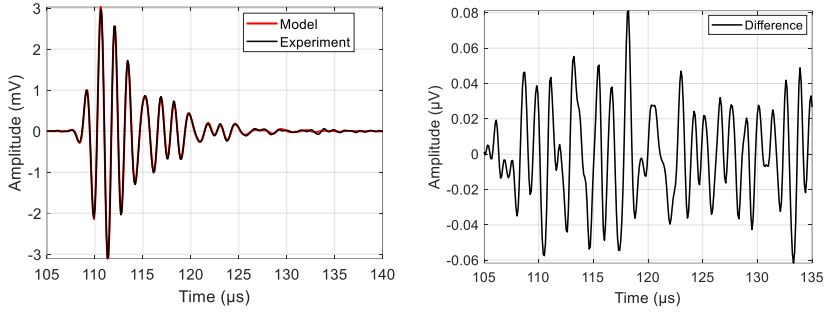


Fig. 40. Time domain TRUS fitting result for sample signal using pulse excitation: model: (left) signal comparison and (right) remainder after model signal subtraction from experimental one

A higher SNR and a broader bandwidth are required to make such a measurement system fit for a broad range of samples. Therefore, SS signals are used in further investigations. The aim is to derive optimal APWP optimization criteria and bandwidth coverage relative to the available -30 dB passband.

3.2. APWP signals

The calibration signal (AC response of the transducer) is obtained using single-pulse excitation signal (Fig. 41), to demonstrate the bandwidth of the transducer. The response is obtained through two 650 kHz air-coupled transducers produced by CSIC, Spain.

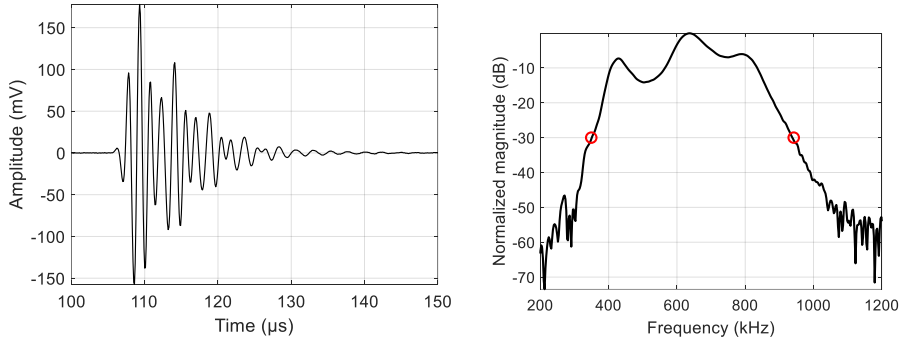


Fig. 41. Received signal after pulse excitation: (left) time domain, (right) normalized in spectral domain (circles: -30 dB bandwidth)

It can be observed that bandwidth is already broad, i.e. a -30 dB bandwidth covers almost 98% of the centre frequency (340-940 kHz). The -30 dB is focused as on this point SNR is sufficient for reliable processing, based on the assumption that beyond -30 dB, a good SNR cannot be achieved; therefore, the frequency range used for further processing will be same as mentioned.

3.2.1. Frequency dependent noise

It is a common assumption that noise is white, thus only signals were used for spectral losses compensation, however Fig. 19, demonstrates the frequency dependence nature of noise, thus it is SNR that is important not only the signal energy that determines the estimation accuracy. The SNR is obtained in frequency domain as the ratio between signal S_{cal} to noise e_n voltage spectral density:

$$SNR = \frac{S_{cal}}{e_n}. \quad (3.2)$$

The NVSD AC response was obtained by measuring the 1000 waveforms of the receiving amplifier output with no excitation signal. The resulting waveforms were converted into frequency domain, averaged power wise, and normalized Eq. 2.9.

The signal S_{cal} , was obtained using LFM with bandwidth coverage of 340 kHz to 940 kHz. Thus, APWP signal derivation was performed based on SNR.

3.2.2. SNR equalization

The compensation function ξ , was obtained using SNR with additional window W to suppress the noise at the edges of the passband region of the transducer i.e. -30 dB, Eq. 2.11.

The SNR based compensation function is shown in Fig. 42.

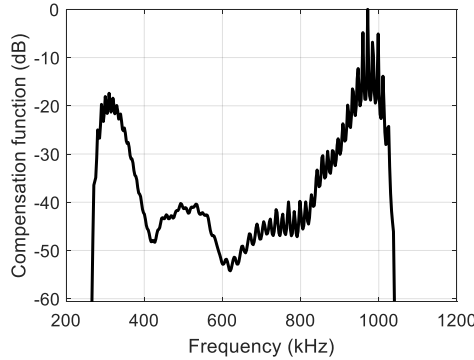


Fig. 42. Compensation function for SNR

Using the compensation function, optimized NLFM signal can be derived, it is important to note that energy of each spectral content is dependent on the rate of the change of instantaneous frequency, therefore, time vs frequency is described using Eq. 2.13, and interpolation in Eq. 2.14, phase is obtained using Eq. 2.15 and unipolar NLFM signals are obtained using Eq. 2.16.

The LFM and NLFM signals have a rectangular envelope (section 1.5.2), which results in large spectral ripples, not recommended for the excitation signals, therefore, further optimization is carried out using APWP signals.

3.2.3. APWP signals for optimal spectrum shaping

APWP signals are based on the arbitrary waveforms which require optimization for adjusting the width and position of the pulses in time. To simplify the electronics complexity, unipolar excitation signals were proposed. Every candidate APWP sequence must be applied in transmission to account for the transducers and electronics' AC response and noise. To speed up the optimization process, the actual measurement was replaced by passing the candidate through a system mathematical model. The transmission model was derived from the response in Fig. 41 and fitted into an 8-th order infinite impulse response (IIR) filter. Noise, coloured by using (Eq. 2.9), was added after passing the candidate sequence through the system model

To optimize the pulse and width position of APWP signals, six different convergence criteria were explored, i) aimed at SNR flatness, ii) aimed at LFM SNR improvement on a linear scale, iii) aimed at LFM SNR improvement on a linear scale with the SNR as an additional weight, iv) aimed at LFM energy improvement on a dB scale, v) aimed at NLFM SNR improvement on a dB scale and vi) aimed at NLFM 1/SNR improvement on a dB scale. Each criterion was focused only on desired frequencies, f_{max} - f_{min} .

The compensation function was developed using SNR_c , the SNR AC response of the current version of APWP signal, calculated using (Eq. 3.1).

Criteria (i), aimed at SNR flatness, is based on standard deviation of SNR_c AC response within the desired passband f_{max} - f_{min} :

$$\zeta_{APWP} = \sqrt{\frac{\int_{f_{min}}^{f_{max}} (|SNR_{Ci}(f)| - |\overline{SNR}_{Ci}(f)|)^2 df}{|\overline{SNR}_{Ci}(f)|}}. \quad (3.3)$$

That is APWP signal is optimized in such way that energy is pushed towards the edges of the passband where losses are high.

Criteria (ii), aimed at SNR improvement on a linear scale. LFM signal (instead of NLFM) was used as initial guess:

$$\zeta_{LFM} = \sum_{f_{min}}^{f_{max}} (|SNR_{LFMi}(f)| - |SNR_{Ci}(f)|). \quad (3.4)$$

That is, APWP signal is optimized in such way that energy is attempted to improve over LFM signal, no matter which frequency.

Criteria (iii) aimed at LFM SNR improvement on a linear scale as per (3.3), but with the SNR as an additional weight. LFM signal (instead of NLFM) was used as initial guess:

$$\zeta_{LFMW} = \sum_{f_{min}}^{f_{max}} \frac{(|SNR_{LFMi}(f)| - |SNR_{Ci}(f)|)}{|SNR_{Ci}(f)|}. \quad (3.5)$$

That is, APWP signal is optimized in such way that energy is attempted to improve over LFM only for frequencies where spectral losses are high.

Criteria (iv) aimed at LFM energy improvement on a dB scale. LFM signal (instead of NLFM) was used as initial guess:

$$\zeta_{LFMdB} = \sum_{f_{min}}^{f_{max}} (20 \cdot \log_{10}|SNR_{LFMi}(f)| - 20 \cdot \log_{10}|SNR_{Ci}(f)|) \quad (3.6)$$

That is, APWP signal is optimized in such way that energy is attempted to improve over LFM, no matter which frequency, same as (Eq. 3.3) only with dB sensitivity.

Criteria (v) aimed at NLFM SNR improvement on a linear scale. NLFM signal was used as initial guess:

$$\zeta_{NLFM} = \sum_{f_{min}}^{f_{max}} (20 \cdot \log_{10}|SNR_{NLFMi}(f)| - 20 \cdot \log_{10}|SNR_{Ci}(f)|) \quad (3.7)$$

That is, APWP signal is optimized in such way that energy is attempted to improve over NLFM signal no matter which frequency.

Criteria (vi) aimed at NLFM 1/SNR improvement on a dB scale. NLFM signal was used as initial guess:

$$\zeta_{1/NLFM} = \sum_{f_{min}}^{f_{max}} (20 \cdot \log_{10}|1/SNR_{NLFMi}(f)| - 20 \cdot \log_{10}|1/SNR_{Ci}(f)|) \quad (3.8)$$

That is, APWP signal is optimized in such way that energy is attempted to degrade over NLFM, no matter which frequency.

The aforementioned criteria consider different aspects of spectra. While (Eq. 3.3) aims to achieve spectral flatness, (Eq. 3.4), (Eq. 3.6) and (Eq. 3.7) allow for any spectral shape change that improves the average SNR, (Eq. 3.5) tries to account for the spectral shape and (Eq. 3.8) is the inverse of the rest, used just to prove that such an approach does not improve the results. The idea that all criteria are used to redistribute energy suggests that this will improve the measurement results. The optimization algorithm pseudocode is shown in Fig. 43, where for each optimization criterion, ζ is substituted by (Eq. 3.3-Eq. 3.8).

```

Initialize: APWP = LFM or NLFM; i = 1;  $\xi_{prev}$  = compute_variability(APWP)
While i <= MaxIterations and improvement > threshold:
  For each pulse n = 1 to N:
    For left_trim l = 0 to 0.99 * Ln:
      For right_trim r = 0 to 0.99 * Ln:
        If l + r > Ln: skip (or set  $\xi_{lr}$  = Inf)
        Resize pulse n: set middle to 1, trim left/right to 0
        Simulate transmit: RX = System_Model(resized_APWP)
         $\xi_{lr}$  = std(SNR_band(RX)) / mean(SNR_band(RX)) # F1 to F2
      End r
    Find min_ $\xi$  over r for this l
  End l
  Find global min_ $\xi(l,r)$ ; Update pulse n with best l_opt, r_opt;  $\xi_n$  = min_ $\xi$ 
End n
 $\xi_i$  = compute_variability(updated_APWP)
 $\Delta\xi_i$  = | $\xi_i$  -  $\xi_{prev}$ | # Or dB: 20*log10( $\xi_i$  /  $\xi_{prev}$ )
If  $\Delta\xi_i$  < threshold: break
 $\xi_{prev}$  =  $\xi_i$ ; i = i + 1
End
Output: Optimized APWP with flattest spectrum

```

Fig. 43. Pseudocode for SNR based spectral compensation

where, N represents the total number of rectangular pulses in the APWP train. Rather than fixing N to a constant for all cases, it was based on the total allowed signal duration. The exact number can vary because the pulse widths and the gaps between them are adjustable, a longer pulse width means fewer pulses will fit, and vice versa. The key point is that N is not a hard-coded value but is essentially determined by how many pulses can be accommodated in the chosen total duration.

The algorithm *trims* the signal into short rectangular pulses and iteratively shortens or shifts each one to flatten the spectrum: the algorithm starts by generating the initial signal (APWP from LFM/NLFM), send it through the channel, and measure flatness (low variability in SNR across frequencies). Next step is pulse by pulse tweaking: for each pulse (detected via zero-crossings), try hundreds of combos: trim left edge by 0 to 99% of its width (shifts it right), trim right edge similarly (shifts left). Simulate each, re-measure flatness, and pick the trim that smooths the spectrum best. After tweaking all pulses, check if flatness improved (e.g., >1 dB). If yes, do another full round (up to 10, but 3-5 often enough). Stop when gains are tiny or max rounds hit. Trimming pulses adds subtle envelope modulation to counter distortions. This brute-force search (nested loops) is computationally heavy but effective for small pulse counts (~total number of pulses derived from signal).

The APWP optimization process improves the excitation signal energy (Fig. 44, left Fig. 46, left), as the variation observed is only ± 1 dB for 100% bandwidth coverage. In comparison to excitation signal energy, the differences between average SNR are more profound.

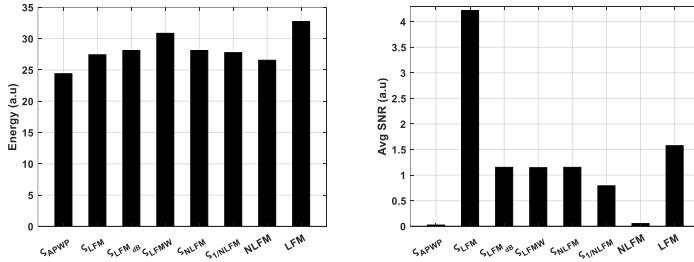


Fig. 44. (right) Energy of excitation signal compared to (left) average SNR of received sample signal for broadest bandwidth coverage, 650 kHz (100% of center frequency)

Although each optimization criterion produces a different excitation code Eq. 3.3-3.8, the energy of these codes are very similar when measured before passing through the physical system (system model). This is because the excitation sequences were designed under equal-energy or equal-duration constraints, meaning that differences arise mainly in how the spectral content is distributed, not in the total input energy. However, this similarity holds only at the transmitter side i.e. without passing through the system response. Once the excitation codes propagate through the real system (Fig. 46) which includes the transducer frequency response, electronics, air attenuation, and sample transmission, the originally similar excitation energies become weighted differently at different frequencies (will be discussed in next

section). Therefore, similar input energy does not translate into equal received SNR, and the differences seen in later figures reflect how well each excitation code interacts with the actual system transfer function.

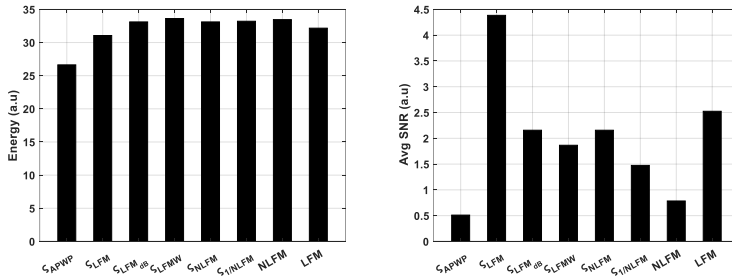


Fig. 45. (right) Energy of excitation signal compared to (left) average SNR of received sample signal for broadest bandwidth coverage, 400 kHz (60% of center frequency)

The excitation code structure plays a dominant role in determining received SNR when wide bandwidth needs to be covered. Even though excitation energies are comparable, the received SNR varies significantly because excitation codes that better match the transducer's usable bandwidth and compensate spectral losses transmit more energy into frequency regions where the system is efficient and noise is low. Rectangular NLFM tends to be more sensitive to spectral nulls, which reduces received SNR despite having similar input energy.

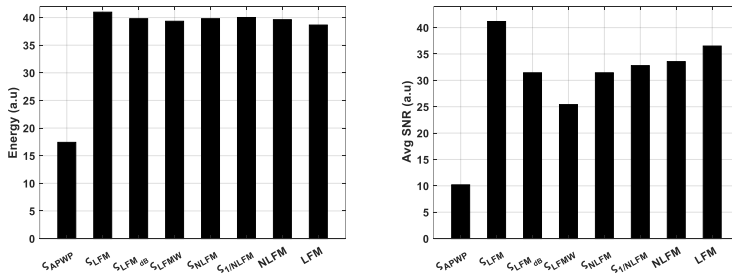


Fig. 46. (right) Energy of excitation signal compared to (left) average SNR of received sample signal for broadest bandwidth coverage, 100 kHz (15% of centre frequency)

Although all excitation codes have similar input energy, their received SNR differs greatly because the system transfer function reshapes each waveform differently. Wideband measurements benefit most from spectrally compensated signals, while narrowband operation reduces the performance gap between methods. Overall, the effectiveness of an excitation code is determined not by its total energy, but by how efficiently that energy is delivered into the system's usable frequency band.

3.3. Experimental setup

The proposed approach was evaluated using thin polycarbonate samples of known thickness 2.05 mm, two 650 kHz transducers in through transmission mode were optimally placed at the fixed distance of 35 mm from each other, as shown in Fig. 47. The electronics for the experimental approach involve programmable voltage excitation pulse, with ability to generate APWP pulse train, and a programmable gain reception amplifier: 8 dB for calibration setup and 50 dB for the sample measurements. Excitation used contain 35 V unipolar sequence that were each of duration 100 μ s long. A simple Bluetooth connection was used to configure the ultrasonic system with the computer for data acquisition, while all subsequent processing—including the inverse solution, optimization, and evaluation was carried out in MATLAB.

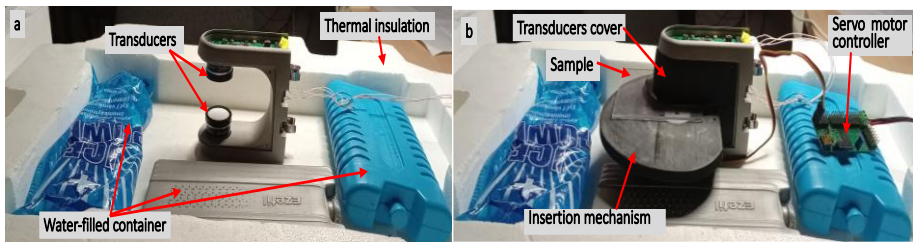


Fig. 47. Experimental measurement setup: (top) equipment view with insertion mechanism removed; (bottom) insertion mechanism view

The aim of this study was to estimate simultaneous parameters of the sample using TRUS, focused was developed on the evaluation of attenuation (α), thickness (h), ultrasound velocity (c) and density (ρ) of the sample. Because temperature, humidity, and pressure strongly affect air-coupled ultrasonic propagation, the sample was placed inside a thermally stabilized chamber containing water-filled containers (Fig. 47). The water mass acts as a passive thermal buffer, reducing short-term temperature fluctuations caused by room convection or equipment heating. This ensures that the air temperature inside the chamber remains nearly constant during the experiment, minimizing changes in sound speed and attenuation.

To reduce the environmental effects experiments were carried out only after 24 hours of wait where setup was exposed to laboratory conditions, to further improve the accuracy, air-parameter compensation described in [43] was adopted. Using servo motor configuration, a sample was inserted and removed for calibration and sample measurements automatically, which were carried out in turn for 100 times.

Three sets of experiments were designed, (i) concentrated around the resonance peak 620 kHz of the sample; (ii) expanding from the resonance peak toward higher frequencies; (iii) expanding from the resonance peak toward lower frequencies. The bandwidth was increased in steps of 25 kHz till the stopband was reached, the spectral combination contained 100 sample-calibration pairs, where each was used for TRUS parameter estimation.

3.4. Results

The transmission response for the polycarbonate sample is presented in Fig. 48. The mean, minimum and maximum values are shown to give an impression of variation caused by noise on magnitude and phase responses.

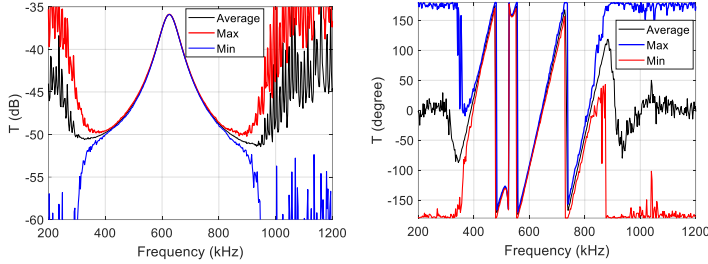


Fig. 48. Transmission response of polycarbonate sample: (left) magnitude, (right) phase

The experimental approach described in previous section was carried out for total of eight signals, LFM, NLFM and six APWP signals (Eq. 3.3- 3.8). Each signal combination produced 100 results for each step of bandwidth variation; three bandwidth placements were evaluated. The results were processed to obtain the mean and then compared against caliper measurements to obtain the bias error.

3.4.1. Spectrum location and bandwidth influence

Three sets of experimental results were focused based on the investigation of bandwidth and its influence on the resonance peak and valleys; therefore, first experiment was focused on the resonance peak where bandwidth was increased in both sides of the resonance peak with step size of 25 kHz till stopband was reached. The second experiment was focused from resonance peak towards the higher frequencies in similar fashion, and last experiment was developed by moving from resonance peak towards the low frequencies. The bias errors for each case were combined and displayed for each parameter, i.e., the ultrasound attenuation α_0 (Fig. 49), velocity c (Fig. 50), density ρ (Fig. 51) and thickness h (Fig. 52) of the sample. Bias errors were calculated by taking the average and comparing it to the actual value ($a = 43.5$ Np/m, $c = 2240$ m/s, $h = 2.045$ mm, $\rho = 1193.6$ kg/m³, [43]) of the polycarbonate sample.

The experimental results for bias errors of attenuation are shown in Fig. 49. Spectral content is concentrated around the resonance peak and expanding both sides (left), expanding from the resonance peak toward higher frequencies (middle) and expanding from the resonance peak toward lower frequencies (right).

The common thing in all three cases is that, when bandwidth was narrow, bias errors were high. Once the resonance peak is covered by spectral content and the bandwidth is midway to the valley (valley is reached at 600 kHz bandwidth for left figure, 300 kHz for centre and right), errors start decreasing. By further expanding the bandwidth, the reduction in bias errors is insignificant.

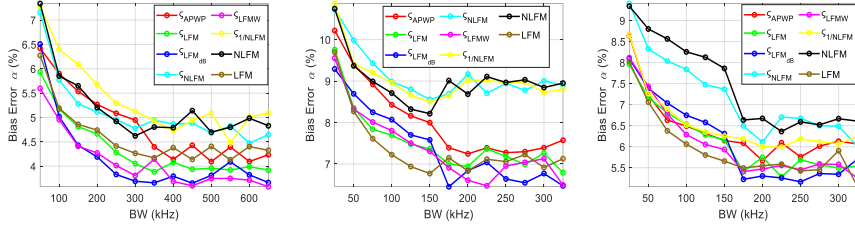


Fig. 49. Attenuation α bias errors for the polycarbonate sample when the excitation coverage is concentrated (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies

The best results achieved are for LFM and LFM-based APWP using criteria ζ_{LFMdb} (where any energy improvement over LFM is favoured). All NLFM-based signals performed the worst. Result can be expected, because attenuation influence is only around the peak. The SNR peak is located at approximately same location as sample resonance peak. Therefore, it is enough to use LFM signal with sufficient spectral coverage (covering peak and half-way to the valley). LFM-derived APWP has better performance, but insignificant.

The experimental results for bias errors of ultrasound velocity estimation using NC-RUS are shown in Fig. 50.

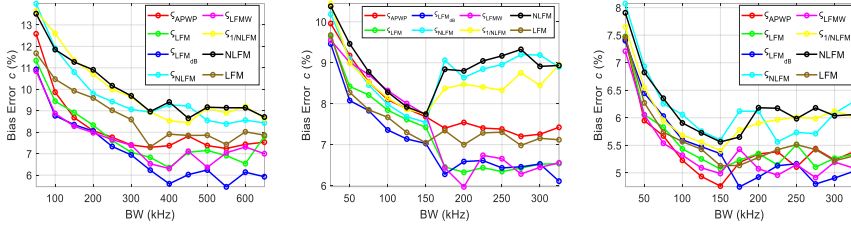


Fig. 50. Ultrasound velocity c bias errors for the polycarbonate sample when the excitation coverage is (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies

The trend is same for the ultrasound velocity case: once resonance peak and half-way to the valley (half-way is at 300 kHz for left and 150 kHz for centre and right) are covered, reduction of bias errors is not that strong. The best results achieved are for LFM-based APWP using criteria ζ_{LFMdb} , ζ_{LFM} (where any energy improvement over LFM is favoured) and ζ_{LFMW} (where energy is attempted to improve over LFM only for frequencies where spectral losses are high). All NLFM-based signals performed the worst. Result can be expected, because velocity influence is mainly around the peak (refer Fig. 51: transmission magnitude). Yet, valley also has some importance, this is the reason why LFM had worse performance (insufficient SNR at valley). Slightly better results are obtained with lower frequencies valley covered (Fig. 51, right).

Bias errors for the density of polycarbonate sample are shown in Fig. 51.

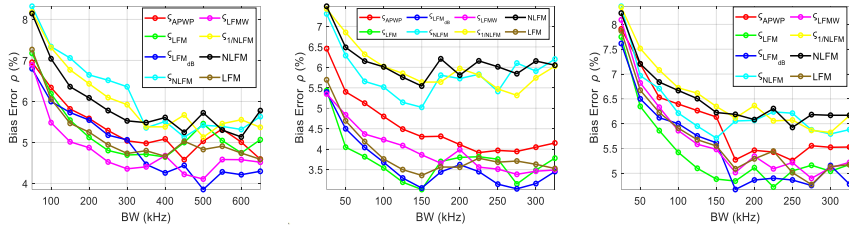


Fig. 51. Density ρ bias errors for the polycarbonate sample when the excitation coverage is (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies

The general trend is same only error reduction continues down to the valley (600 kHz for left and 300 kHz for centre and right), reduction of bias errors beyond this point is not significant. Clear winner can be seen: LFM-based APWP using criteria ζ_{LFMdB} (any energy improvement over LFM is favoured). All NLFM-based signals performed the worst. Result can be explained: despite both peak and valley are influenced by density, density estimation is based on impedance, therefore, it is related to velocity (influence around the peak). Yet, the resonance valley also has some importance, therefore LFM had worse performance.

Bias errors thickness measurement are shown in Fig. 52.

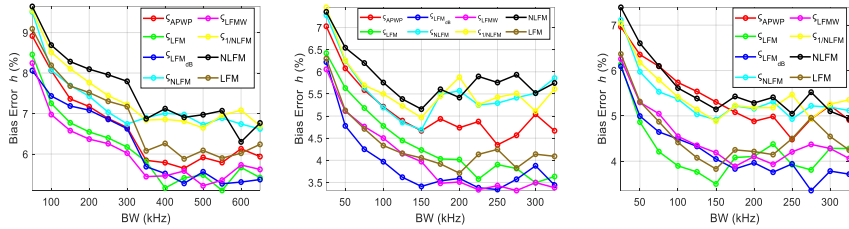


Fig. 52. Thickness h bias errors for the polycarbonate sample when the excitation coverage is (left) around the resonance peak, (middle) expanding from the resonance peak toward higher frequencies and (right) expanding from the resonance peak toward lower frequencies

Similar trend is observed: bias error reduction until the valley is reached, but it is interesting to point out that single side valley coverage produces better results (errors down to 3.5%) than covering both valleys (5.5% errors). Results are better if higher frequencies valley (situated right from the peak) is covered. The best results are achieved for LFM-based APWP using criteria ζ_{LFMdB} (where any energy improvement over LFM is favoured) and ζ_{LFMW} (where energy is attempted to improve over LFM only for frequencies where spectral losses are high), both are dB scale. Result can be expected, because thickness influence is mainly around the peak. Yet, broad coverage also has some importance, this is the reason why LFM had worse performance (insufficient SNR away from peak).

Results above can seem unexpected because criteria ζ_{LFMW} is producing the spectrum that is most flat (see Fig 53). Then results for it should be the best, but Fig.53 explains why: optimization is “punished” even for better SNR (note the arrows) since this criteria aims flatness, ignoring gain in SNR.

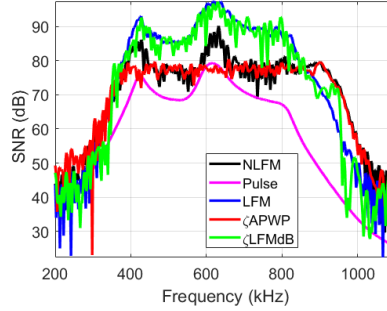


Fig. 53. Spectral response of compensation SNR using pulse, NLFM, LFM, ζ_{APWP} , and ζ_{LFMdB} . (pulse signal is not compensated)

To answer the question which part of the spectrum energy should be focused, assumption could be to focus energy on resonance valley as peak already contains sufficient energy. However, through investigations above, it is observed that both resonance peak and valley are important, and once resonance peak and one valley is covered by the bandwidth, bias errors are reduced and further increasing the bandwidth does not improve a lot. Yet, essential conclusion is that further bandwidth increase does not increase the errors. Then, final conclusion can be drawn: wide bandwidth is important (but within 10-20 dB of loss compensation), but any signal modification that improves the SNR within the selected bandwidth reduces the errors. Therefore, LFM-derived APWP using ζ_{LFMdB} criterion provides the best results.

3.5. Conclusions

Evaluation of the spectral coverage’s effect on measurement errors have demonstrated that it is enough if spectral content covers the sample resonance peak and one valley. The lowest errors for density (3%), and thickness (3.5%) are obtained in the case when spectral content covers the resonance valley which is on the high frequencies part. When the spectral content covers lower frequencies valley, the best result is for velocity (5%). Including both valleys result in the minimal error for attenuation (3.8%). Compromise, satisfying all parameters is when the entire -20 dB passband is covered, no discernible performance drop is seen.

It was demonstrated that APWP signals offer better results than the signal that they are derived from. LFM-derived APWP using ζ_{LFMdB} criterion provides the best results. LFM signals with wide spectral coverage can also be used with few percent errors increase.

4. SOLUTION FOR LONG SIGNALS OVERLAPPING

The accuracy of signal interpretation in ultrasonic measurements relies on the temporal separability of successive reflections [120]. In previous chapters, discussion is developed around wave propagation, measurement techniques, excitation signal optimization and transducers development, however, all these focuses on the compensation of instrumental and environmental constraints. The persistent challenge that undermines the accuracy and resolution particularly in TRUS, is signal overlapping [121]. The overlap in signals is primarily categorized into three domains: physical, transducer based and signal related. Physical constraints appear due to short propagation distance and multiple reflections of a thin sample. Transducer based limitation happens due to finite ring-down duration and insufficient damping, which results in continuous oscillations even after the end of excitation signal, making early reflections. Lastly, constraints related to the excitation signal arise from narrowband transducers, long coded signals with poor temporal localization blur the reflection sequence, while even wideband pulses can generate interference if phase distortions occur along the propagation path [122]. The combined result is an inseparable mixture of overlapping reflections, in which neither direct deconvolution nor standard correlation analysis can isolate the individual contributions reliably.

4.1. Overlapping in long duration excitation signals

The AC response of LFM signal with two visible reflections are shown in Fig. 54, the important thing to consider here is that if the reflections are distinguishable, then gating (tapering) is possible as seen in Fig. 54 (left and middle). However when the excitation duration is long i.e. $250\text{ }\mu\text{s}$ (Fig. 54 right), conventional gating is impossible, also simple matched filtering is also difficult in such cases.

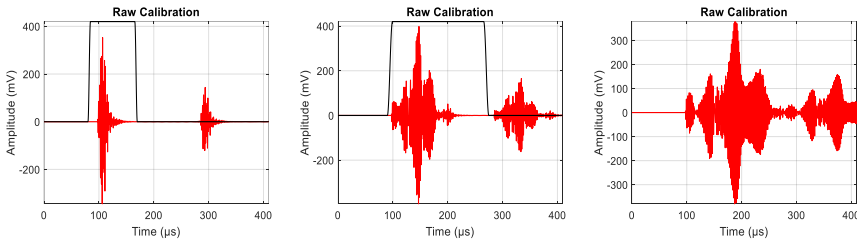


Fig. 54. Calibration signal (AC response of transducer) of (left) $30\text{ }\mu\text{s}$, (middle) $100\text{ }\mu\text{s}$ and (right) $250\text{ }\mu\text{s}$ long LFM excitation signal

The sample measurement depicts different story, as seen in Fig. 55, where even the shortest ($30\text{ }\mu\text{s}$) excitation signals contain distortions that cannot be effectively removed using gating. The case becomes worse for longer duration ($100\text{ }\mu\text{s}$ or $250\text{ }\mu\text{s}$), where multiple reflections are overlapping so much that it is impossible to develop understanding of source reflections (Fig. 54). Due to this very reason, conventional gating is not feasible for the transmission response estimation using inverse problems which is essential part of TRUS application.

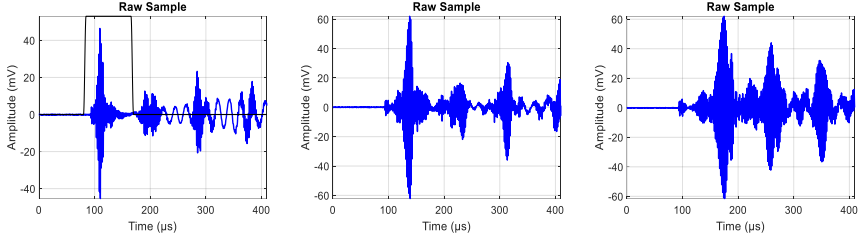


Fig. 55. Sample signal (sample inserted between transducers in through transmission of air-coupled ultrasound) of LFM with (left) 30 μs , (middle) 100 μs , (right) 250 μs excitation signal duration

The transmission response of different duration signals are shown in Fig. 56, where distortions can be visibly seen in the transmission response of long duration signal (Fig. 56, right). As overlapping reflection disrupts the transmission so much that resonance peak and valley are not detectable, rendering all efforts for inverse problem of TRUS parameter estimation useless.

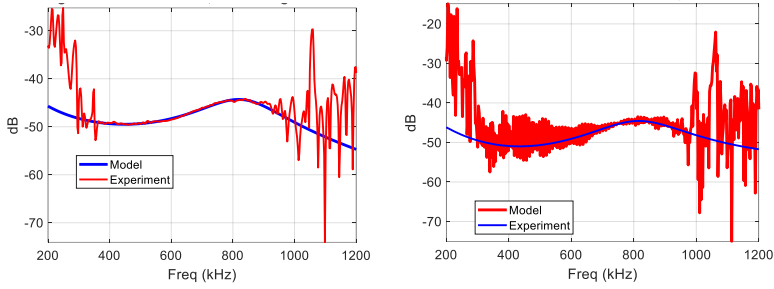


Fig. 56. Transmission response of (left) *vitis vinifera* (grape plant leaf) when reflections are separate, (right) when reflections are overlapping

The total energy E of ultrasonic signal with voltage excitation $s(t)$ in frequency domain can be determined as,

$$E = \int_{-\infty}^{\infty} |s(f)|^2 df. \quad (4.1)$$

If the signal has an approximately flat spectrum with bandwidth B , (i.e. idealized rectangular or chirp spectrum);, then through Parseval's theorem [123],

$$|s(f)|^2 \approx \frac{A^2}{B} \text{ for } f_0 - \frac{B}{2} < f < f_0 + \frac{B}{2}. \quad (4.2)$$

Integrating over B :

$$E \propto A^2 \cdot B. \quad (4.3)$$

The relationship of total energy is therefore, directly proportional to the square of the amplitude and bandwidth. To increase the transmitted energy, the excitation amplitude is typically increased, however it cannot be increased inevitably due to limitations of linear amplifiers, and DACs. Moreover, the transmitted ultrasonic energy is proportional to the square of the excitation amplitude, doubling the energy

requires increasing the amplitude by a factor of $\sqrt{2}$, not four. The other way around this problem is to increase the bandwidth (duration) of the signal as it is directly proportional. Chapters III and IV, focused on development of long duration signals, however as discussed earlier, signals cannot be increased in duration due to limitation of transducer distance placement in air-coupled through transmission mode. Therefore, there is need for the solution where the overlapping problem is solved.

4.2. Deep neural network approach for overlapping solutions

Air-coupled ultrasound in through transmission, requires calibration and sample measurements, taking ratio in frequency domain results in the transmission response of the sample measurements [36,42]. Conventional approach is to use short pulse signal for the excitation of transducer; it provides excellent temporal resolution but lacks in bandwidth and SNR (discussed in Chapter III and IV). Spread spectrum signals offer the ability to provide long duration waveforms in which the signal energy is distributed over a broad range of frequencies across time [36,42,50], LFM signals are one such spread spectrum signals (Fig. 57).

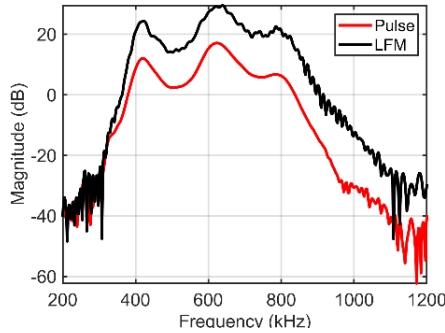


Fig. 57. Spectral response comparison between LFM and pulse excitation signals

As in air-coupled configuration, transducers cannot be much far distance placed from each other; therefore, measurement could be taken for the short duration (10 μ s till 100 μ s in calibration measurements) without any overlapping reflections, for the longer duration i.e. 100 μ s onwards, it is extremely difficult and complex to isolate the individual reflections. The time and frequency domain representation of 50 μ s vs 250 μ s is shown in Fig. 58, where longer duration signals results in complex overlaps that cannot be separated using conventional methods.

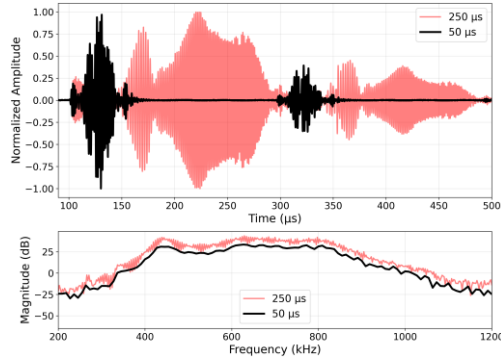


Fig. 58. (top) Temporal response, (bottom) spectral response of 50 μs chirp compared with 200 μs chirp

Simple signal overlapping is manually solved using matched filter [124], manual gating [125], or frequency domain processing. Manual gating requires selection of time windows to isolate the reflections, in case of such overlap in reflections that there is no distinguish between reflection, match filtering become quite complex. Matched filtering or inverse filtering provides improved temporal resolution based on the assumption that the system is linear, and the impulse response is known, which is almost impractical in air-coupled systems.

Deep neural networks (DNN) are gaining weight in signal processing especially in overlapping signals or complex signals processing [126,127]; however, the conventional approach is either segmentation or classification of signals. In [128], a novel DNN based experimental air-coupled ultrasound method was proposed which unlike classification or segmentation, the proposed method have the ability to train on experimentally obtained data but, instead of only classification, predict for the data which was not even part of the training.

The experimental setup was developed with 650 kHz air-coupled transducers, for different durations of LFM (10 μs to 350 μs). In [128], a residual DNN based approach was proposed, with goal of solving the overlapping signals in long-duration signals. The model training was achieved for 10 μs to 50 μs short duration signals where the reflections are distinctive from each other without any overlapping. Only first reflection, i.e. c_1 was gated for training, the excitation voltage on the transducers were measured and considered as the input to the model, the output of first reflection of calibration signal for the DNN model. The target function was developed as L_2 norm of both frequency and time domain of predicted signals and experimental signals. This ensures that both temporal and spectral features are part of the training with aim that model is trained in such way that once it trains on short duration signals, it can predict from the long duration overlapping signal a non-overlapping clear reflection. The modelling was done using U-shaped network architecture (UNET) model that already exist in literature; however this existing model was modified for the specific task of signal overlapping solution using 1D convolutions and residual channels.

4.2.1. Res-UNet1D based DNN model

Air-coupled transducers in through transmission were used for data acquisition. Data collected and stored in mat format where 100 repetitions were measured for each duration (i.e. 10 μ s till 350 μ s) for a better statistical analysis of a diverse training set. Data was partitioned between training and testing: training set was 10 μ s till 50 μ s, and rest of all other signals was reserved for testing of proposed approach. The methodology is presented in Fig. 59.

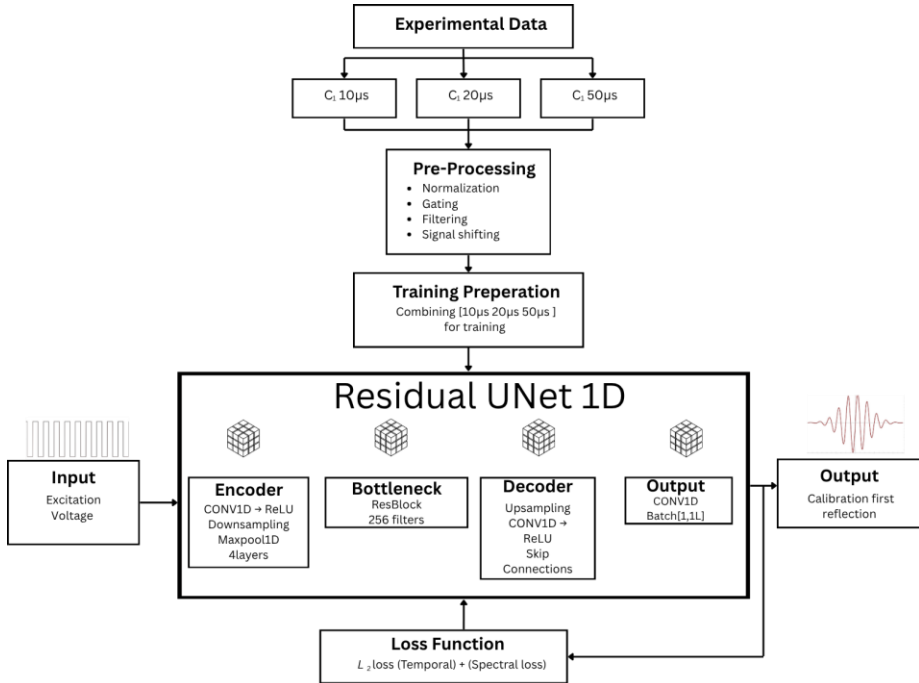


Fig. 59. Res-UNet1D based methodology flow diagram

The rectangular chirp was used for the excitation signals, both excitation voltage and calibration signals were normalized before processing to ensure smooth training process. To remove unwanted noise influence, bandpass filter of 0.15 MHz-1.5 MHz was applied to the signals. The UNET model was developed from scratch (although variants are available, however new variant was developed) with residual UNet1D (Res-UNet1D) designed for residual tasks involving air-coupled ultrasound measurements, which are 1-dimensional (1D) (A-scans) in time domain. The residual block consists of two 1D convolutional layers, followed by batch normalization and activated leaky rectified linear unit (LReLU) functions. Four staged encoding was done where each stage contained dual residual blocks with progressively increasing dilation (1,2,4,8). Such dilation was adopted to ensure that model effectively captures complex temporal features on different dataset durations. The detailed architecture with tensor shape (i.e. (1, 1, 6000), (1, 2, 1500) etc.) for the convolutional neural network is shown in Fig. 60. Temporal compression in the encoder reduces redundant

structure and enables disentanglement in the bottleneck. The decoder reconstructs the cleaned waveform with phase and amplitude consistency. Unlike conventional max-pool architectures, stride-based down sampling preserves continuity of the signal.

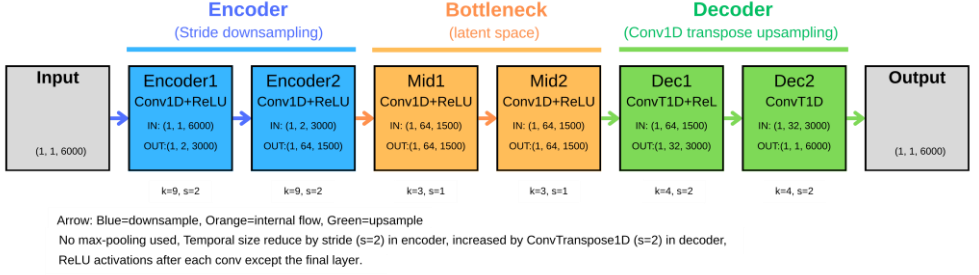


Fig. 60. Architecture of the 1D convolutional autoencoder model

The bottleneck segment increases the feature extraction process with dual residual blocks with high dilation rate of 16. The decoder mirrors the encoder path with slight up-gradation of using ConvTranspose1D for up-sampling. Output is generated in a sequence of 1 x 1 convolutions projecting to a single channel output.

The model fitting result of training is shown in Fig. 61, where there is almost perfect resemblance between both temporal and spectral response.

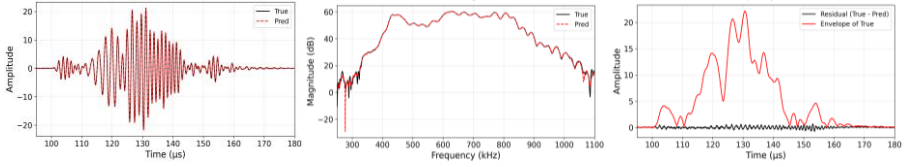


Fig. 61. (left) Temporal, (center) spectral and (right) residual response of 50 μ s that was part of the training

The total loss function (L_{total}) is defined by fitting in both time domain (L_{time}) and frequency domain (L_{freq}),

$$L_{\text{total}} = L_{\text{time}} + \alpha \cdot L_{\text{freq}}, \quad (4.4)$$

$$L_{\text{time}} = \frac{1}{N} \sum_{n=1}^N (y_n - \hat{y}_n)^2, \quad (4.5)$$

$$L_{\text{freq}} = \frac{1}{K} \sum_{k=1}^K w_k (|F\{y\}_k| - |F\{\hat{y}\}_k|)^2, \quad (4.6)$$

where, $\alpha = 0.1$ is trade-off weight (frequency loss scaling factor) for the total loss function that was set at such a value to obtain optimal or stable performance across our dataset after several validation experiments. y_n and $\hat{y}_n$ are the experimental and predicted signals respectively, N is the total number of samples, $F\{y\}_k$ is discrete Fourier transform of the y at frequency K , and w_k is weighting coefficient for the frequency bin k and w_k . The weight of w_k is set to 1 for frequency bins within the range of [200 kHz, 1200 kHz] frequencies which are around the passband region of 650 kHz air-coupled transducers, and 0.5 for every other frequency bins.

4.2.2. Res-UNet1D model results for long duration signals

The Res-UNet1D successfully demonstrates the learning process for training data (Fig. 61); however real question is about how well model predicts in case of unseen duration. As long duration signals were not part of the training so it will be a unique approach of producing results, the validation was tested for all the datasets. The fitting results for 250 μs LFM signal are shown in Fig. 62, compared with the residual (subtraction of predicted signals from actual in time domain) of 50 μs , the 250 μs is quite large, however, as this was not part of the training and model learns from the shorter duration to predict long duration, this is quite acceptable as noise presence cannot be ignored.

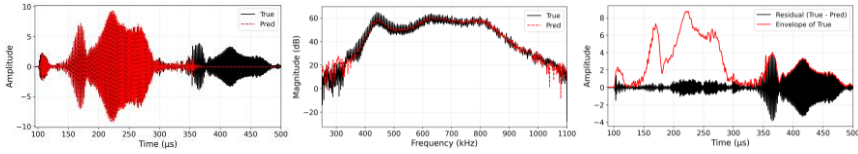


Fig. 62. (left) Temporal, (centre) spectral and (right) residual response of 250 μs that was part of the training

The fitting results for the 350 μs (longest from the dataset), are shown in Fig. 63, the visual results are also acceptable with minor losses of the modelling results. The residual part indicates that learning was successful for the first reflection region. while comparison with actual signal validates the proposed approach. Although for the long-duration signals there are some residuals, however, such little compensation needed to achieve modelling.

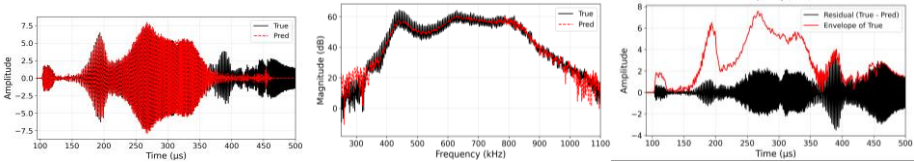


Fig. 63. (left) Temporal, (centre) spectral and (right) residual response of 350 μs that was part of the training

The remainder energy represents the time-domain energy of the residual between the measured and simulated signals. It is calculated using Eq. 4.5 for every signal duration in the dataset, as summarized in Table. 5.

Table 5. Remainder energy and correlation of each duration signals

LFM duration	Remainder energy	Correlation (r)
10 μs	45 $\pm$ 2 dB	0.90 $\pm$ 0.03
20 μs	45 $\pm$ 2 dB	0.90 $\pm$ 0.03
50 μs	45 $\pm$ 2 dB	0.90 $\pm$ 0.03
100 μs	42 $\pm$ 2 dB	0.85 $\pm$ 0.03
150 μs	40 $\pm$ 2 dB	0.82 $\pm$ 0.03

200 μ s	39 $\pm$ 2 dB	0.78 $\pm$ 0.03
250 μ s	35 $\pm$ 2 dB	0.75 $\pm$ 0.03
300 μ s	32 $\pm$ 2 dB	0.74 $\pm$ 0.03
350 μ s	28 $\pm$ 2 dB	0.72 $\pm$ 0.03

Res-UNet1D based deep learning approach was proposed for the long-duration reflection overlapping in chirp signals used for air-coupled ultrasound. The results emphasize the importance of a deep learning architecture for modelling and predicting signals with significant overlap. The pre-processing, gating, scaling, normalization, and signal alignment ensured robust modelling of complex ultrasound reflections for experimental data. The close match between predicted and true signals in both the time and frequency domains highlights the model's capability to generalize across different excitation conditions and signal complexities

This study is only concluded for calibration signal modelling, which is less complex compared to sample or transmission response, however, since this approach has never been tried before, its successful application so far is limited to calibration modelling. But in the future, the idea is to extend this methodology for sample measurement, moreover the robustness of this approach is only tested for same type of signals i.e. LFM only, it would be interesting to test the fitting results if model is trained only for LFM signals but what would be the output if validated for different type like pulse or NLFM signals.

4.3. Deconvolution for Long Spread Spectrum Signals

Ultrasound face constant trade-off between signal energy, resolution and bandwidth, short pulse provides sharp temporal resolution, but lacks in energy and bandwidth, a long toneburst carries more energy, but suffers from bandwidth and resolution, Fig. 64. As discussed in previous chapters, energy is the most important for air-coupled ultrasound measurements due to attenuation and transmission losses. SNR can be improved by increasing the pulse amplitude, but there are physical limits for maximum excitation voltage, including nonlinearity in air [129]. Spread spectrum signals were proposed as solution to such problems [90,130-132], the advantage of using SS signals is that they can be suppressed using pulse compression (match filter or cross-correlation) to increase the amplitude of the signal, which result in increased SNR [132].

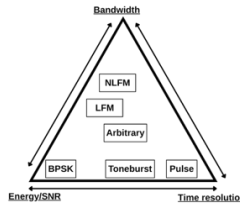


Fig. 64. The trade-off between resolution, energy, and bandwidth in signal design

The principle of pulse compression strategy is to allow transmitters to emit long-duration, high energy signals, while still achieving a resolution of a short pulse upon reception. This is achieved not by shortening the transmitted pulse, but rather by encoding it through phase or frequency modulation. This results in reflections that can later be compressed in time by matched filtering or correlation processing.

Mathematically, pulse compression exploits autocorrelation property of the modulated pulse. If the transmitted waveform $s(t)$ and $h(t) = s^*(-t)$ is its matched filter $h(t)$, then the convolution $y(t)$ is achieved as,

$$y(t) = s(t) * h(t), \quad (4.7)$$

The convolution results in sharp main lobe (the compressed pulse), with ideally minimal sidelobes. The ratio between the transmitted pulse duration and the compressed width defines the compression ratio, directly influencing the achievable SNR gain. The autocorrelation/match filter of pulse, LFM and NLFM signals are shown in Fig. 65. The pulse signal exhibits wide mainlobe with no sidelobes, while LFM shows near-range sidelobes and NLFM suppresses the near-range sidelobes.

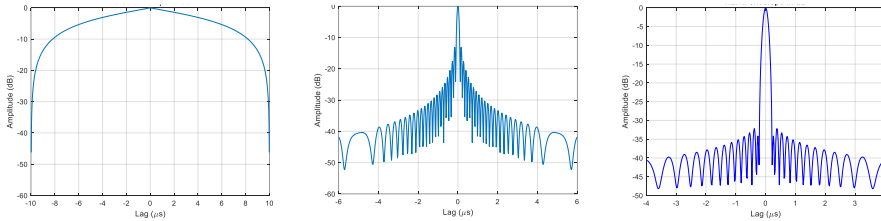


Fig. 65. Autocorrelation of (left) pulse, (centre) LFM and (right) NLFM signals

A wide variety of coding techniques have been developed LFM or NLFM are the most common in ultrasound and radar, providing robustness against Doppler effects and easy implementation. Others, such as phase-coded sequences (e.g., Barker or polyphase codes), are used when phase agility and spectral control are desired. Each method encodes the pulse differently but shares the same decoding principle: maximize the correlation peak at the correct time delay, suppress all others.

The problem with SS signals for pulse compression is that they produce sidelobes, which makes it difficult to maximize the correlation peak. In imaging applications this is not an issue, as it can be solved using signal envelope tapering in time [134] or mismatch filter applications [133]. However, envelope tapering cannot be applied to unipolar or bipolar coded excitation signals, and mismatch filter does not provide desired results. Essential problem with pulse compression in sidelobes is that neighbouring reflections interfere and spectral content becomes disoriented. One approach is to manually gate the compressed signals; however this results in part of spectral information lost, which becomes useless for spectral analysis. Lastly, reflections that are separated by less than the axial resolution (which depends on the signal bandwidth and is ultimately limited by the transducer's bandwidth) cannot be resolved when mismatch filtering is used. Here, mismatch filtering refers to filtering the received signal with a template that does not perfectly match the transmitted

waveform, which reduces the system's ability to distinguish closely spaced reflections.

NC-RUS is a viable solution for reverberations inside the material thickness overlaps [37,96], which results in constructive and destructive interference known as thickness resonances. Ultrasound propagation time can be deduced by studying these resonances, with known velocity thickness can be determined. The transmission spectra (ratio of sample and calibration in frequency domain) produces thickness resonances specific to the sample material thickness and velocity (discussed in chapter 2.2), there is a signal shift in time between Calibration and Sample measurements which is also related to related to sample thickness and ultrasound velocity. These two independent values allow simultaneous estimation of sample thickness, density, ultrasound velocity, and attenuation by solving the inverse problem using the sample transmission model [116].

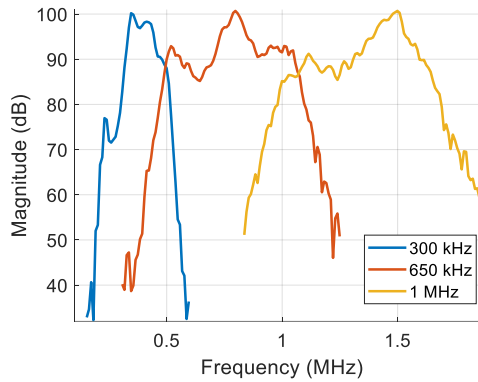


Fig. 66. Calibration signals AC response of three different centre frequency transducers i.e. 300 kHz, 650 kHz and 1 MHz

The application of NC-RUS is well recognized because of SNR and bandwidth requirements; high SNR is requirement for signal transmission losses and the attenuation and density estimation through the inverse solution also depends on how large the SNR is in the resonance valley. As discussed, the position of resonance peak depends on the speed of sound and thickness of sample, if the resonance peak is not within the passband region of transducer, estimation quality degrades. Lastly, as attenuation is highly dependent on frequency, therefore it is also affected by bandwidth. The common way of doing NC-RUS is combining several different centre frequency transducers to get a wideband measurement with single pulse excitation [32,96,], as shown in Fig. 66.

In comparison, SS signals are proven to be more energy efficient and wide bandwidth signals, however in all previous investigation, the duration of SS signals used were either shorter duration or comparable with reverberation time. For air-coupled investigation, this duration was only 50 μ s, however long duration SS signals are necessary to accumulate sufficient energy. Also, spectrum programming flexibility of APWP signal [36,42,50] depends on signal length: the longer is the

signal, the more flexible and more accurate is the spectral shaping. Solution can be pulse compression, however, NC-RUS relies on the accuracy of spectral content; therefore, if manual gating distorts the spectral content, its accuracy is highly affected [135,136].

The idea proposed here is to use inverse filter for pulse compression, which is intermediate between wiener filter and inverse filter [55-57]. The filter was designed using SS coded signals, additional bandpass filter with bandwidth corresponding to the passband of transducer was applied to limit noises. Such approach provides compression that corresponds to the ideal delta pulse. This result in improved SNR while the received signal is like very broadband pulse excitation. Weiner filter [137] can be used to turn the filter into matched for regions where SNR is low. Inverse filter was used only for pulse compression [117]; the compressed signals were gated manually for each reflection, and signal was decompressed back and used for fitting.

4.3.1. Problem analysis

Simulations and experiments were conducted using 2 mm thick polycarbonate plate measured using 650 kHz centre frequency air-coupled transducers through transmission mode. Single wideband 100 ns duration rectangular pulse was used as excitation signal; Calibration was performed using a 10 V excitation, while the samples were measured using a 20 V excitation. The transducers used were very sensitive and broadband, as shown in Fig. 67.

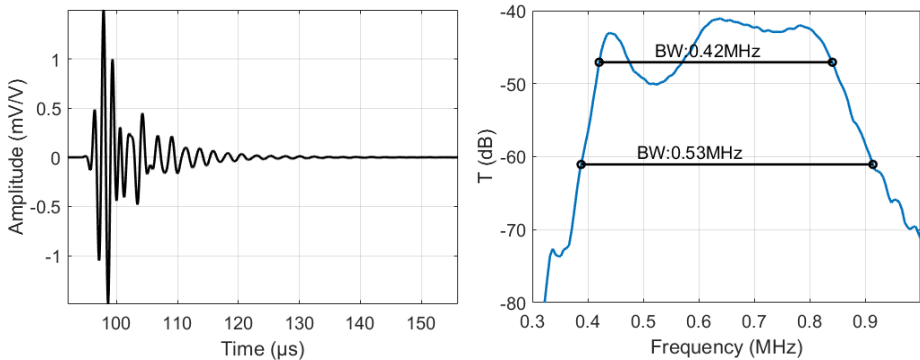


Fig. 67. Calibration signal (left) time domain and (right) frequency domain AC response in case of wideband 100 ns pulse excitation. Bandwidth at -6 dB and -20 dB

Signal was converted to transducer output (preamplifier gain accounted) and normalized to excitation voltage. Transmission response in Fig. 67 (right) is the ratio of received to excitation signal amplitudes. Relative transmission bandwidth is 65% at -6 dB and 81% at -20 dB level. The main problem is evident (Fig. 67, left), where additional reflections can be seen in the signal tail present due to matching layers. Any coded excitation (SS signal) will be a convolution with the response.

4.3.1.1. Pulse compression using standard approach

The calibration signal with 50 μs LFM from 250 kHz-1.1 MHz range rectangular bipolar signals with 10 V (Fig. 68, left), the corresponding pulse compression (Fig. 68, right), using matched filter:

$$C_{MF}(\omega) = C(\omega) \cdot \text{conj}(G(\omega)), \quad (4.8)$$

$$S_{MF}(\omega) = S(\omega) \cdot \text{conj}(G(\omega)), \quad (4.9)$$

where, $C(\omega)$ is the calibration, $S(\omega)$ is the sample and $G(\omega)$ are the excitation signals in frequency domain.

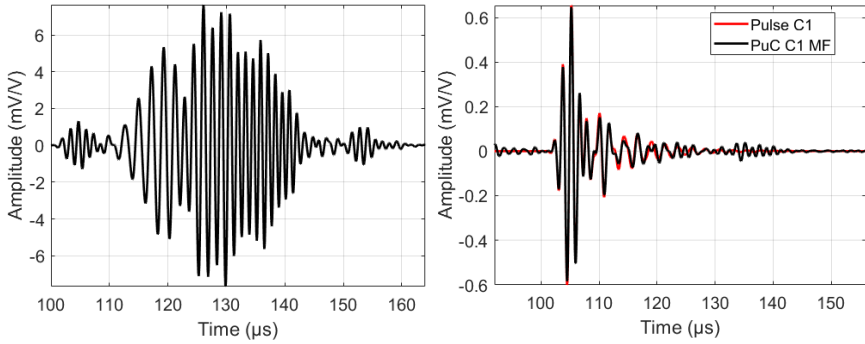


Fig. 68. (left) Calibration signal using LFM and (right) corresponding pulse compression using matched filter, red comparison to pulse (scaled)

Matched filtering results in additional pulse, which are addressed as correlation sidelobes [133]. The result of pulse excitation is shown in Fig. 69, where situation is even worse as correlation sidelobes are now on both sides of the compressed pulse (mainlobe).

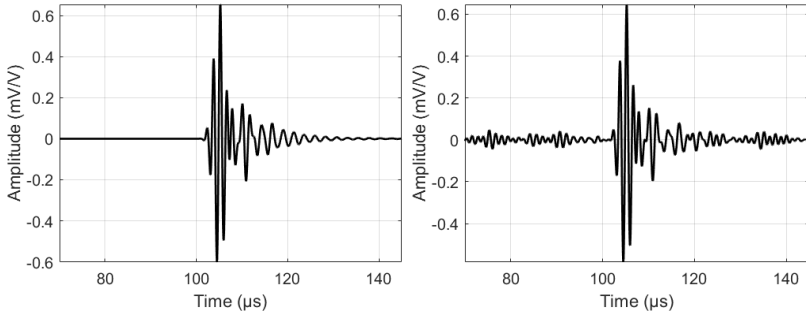


Fig. 69. (left) Comparison of pulse excitation and (right) MF compressed LFM. Compressed signals were normalized

Mismatched filter using minimum four-term Blackman Harris window $BH(\omega)$:

$$C_{MMF}(\omega) = BH(\omega) \cdot C(\omega) \cdot \text{conj}(G(\omega)), \quad (4.10)$$

$$S_{MMF}(\omega) = BH(\omega) \cdot S(\omega) \cdot \text{conj}(G(\omega)), \quad (4.11)$$

The use of mismatched filtering also fails to resolve the problem (Fig. 70), as there are many more distortions compared to pulse excitation.

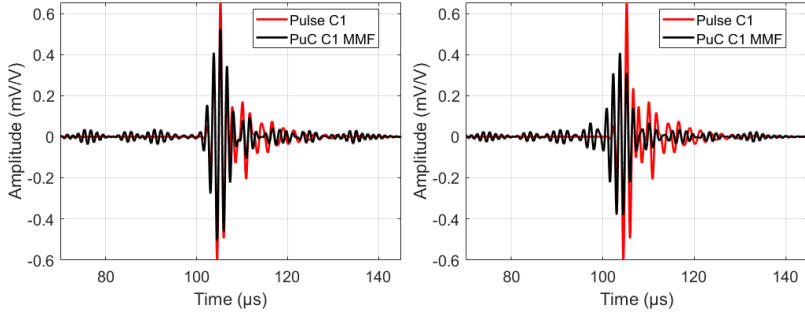


Fig. 70. Mismatched, Blackman-Harris filter-based, pulse compression when reference signal is: (left) excitation code and (right) calibration signal

The choice of the reference signal does not change the output, as mismatched filter aims to reduce only the near range sidelobes at the expense of central lobe broadening [118]. Due to more reverberations in the sample measurements, the results are even worse (Fig. 71). The sample is 2 mm thick polycarbonate ($c=2220$ m/s), round-trip travel time is $1.8 \mu\text{s}$. Signal is much smaller ($\mu\text{V/V}$ vs. mV/V for calibration) due to the reflection losses from sample (impedance mismatch is significant).

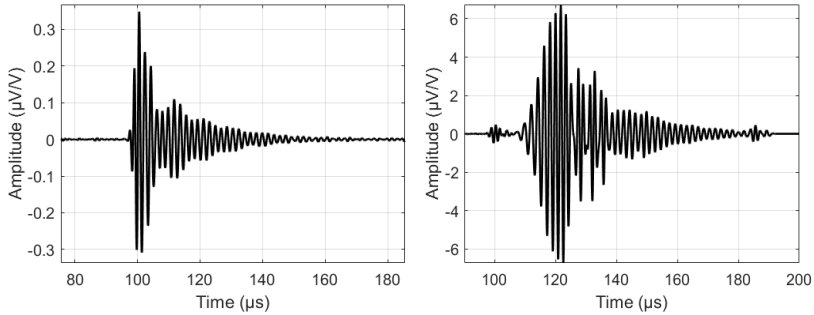


Fig. 71. Sample measurements using (left) 100 ns pulse excitation and (right) $50 \mu\text{s}$ LFM excitation

Reverberations inside the sample reflect valuable information about the sample, which is used for the parameter's estimation. Fig. 72, presents the spectral response of calibration, sample and corresponding transmission response. NVSD estimated using described approach [36], is also presented for visual SNR estimation. Sample

transmission response is based on the experimentally measured (Fig. 72, right, red) and estimated inverse solution (Fig. 72, right, black).

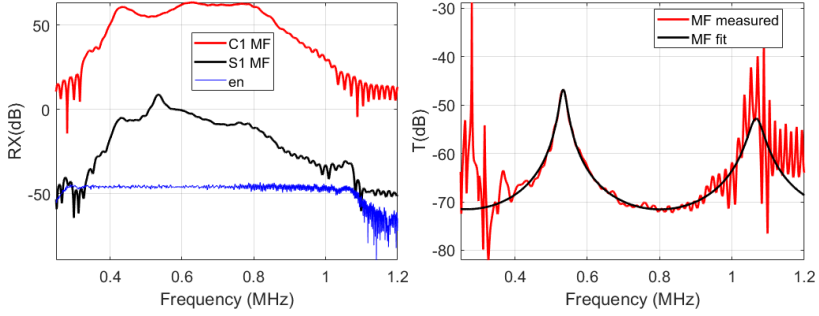


Fig. 72. (left) Spectra of calibration (red), sample (black) and noise (blue) signals vs. (right) measured (red) and fitted (black) sample transmission response

The difference between matched and mismatched filtering in time-domain is illustrated Fig. 73.

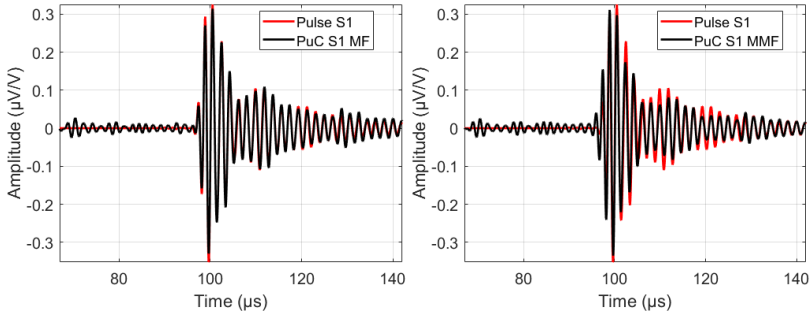


Fig. 73. Comparison of sample signal pulse compression result using (left) MF and (right) MMF. Signals normalized to pulse

The distance between transducers is 35 mm, which results in 100 μs reverberation time. With sample inserted propagation distance reduces to 16 mm, or 47 μs reverberation time. Then 50 μs duration signals will overlap (Fig. 74).

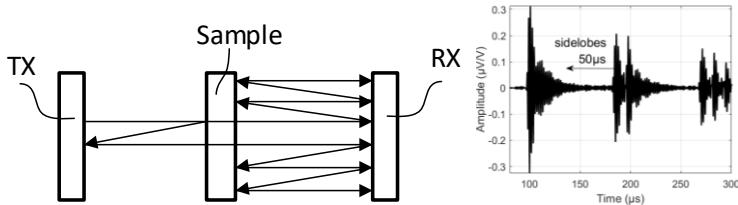


Fig. 74. Sample signal contains overlapping reflections: (left) explanation of propagation; (right) pulse compression result using MF

To achieve the processable spectral response, additional reflections which are visible at span of 50 μs need to be gated out. The inverse solution is obtained in frequency domain; therefore, gate should start from 88 ms till 180 ms, however, if the sample is not inserted exactly in the middle, the second reflection is doubled due to reflections from both transducers. The overlapping of second reflection is at 130 μs . If the gate is set at position where sample reverberation is not decayed sufficiently, a spectrum convolution with *sinc* function will occur. Sample parameters estimation will develop bias errors.

The time domain fitting proposed in [36,42,50] can only be applied in case of short duration i.e 50 μs signals. For high SNR long duration signals are preferred, higher the SNR lower the estimation errors and wider frequency range. For long duration signals (>100 μs) overlapping sidelobes will cover whole signal, and fitting will contain sidelobes, which results in bias errors. For very long duration (300 μs), the sidelobes from third and fourth reflections will also appear in overlapping, inverse solution can be estimated; however, bias errors can increase.

4.3.1.2. Pulse compression using inverse filter

The solution proposed is to use inverse filter for pulse compression:

$$C_{IF}(\omega) = W(\omega) \cdot C(\omega)/G(\omega), \quad (4.12)$$

$$S_{IF}(\omega) = W(\omega) \cdot S(\omega)/G(\omega). \quad (4.13)$$

Similar to previous expressions, $G(\omega)$ is the excitation code or voltage, filter $W(\omega)$, is optional intended for the noise reductions in low SNR regions. 4-th order Butterworth bandpass filter (termed as inverse filter) and Wiener-like filter were used. Wiener-like filter is termed as Wiener filter, defined as,

$$W(\omega) = \bar{C}(\omega)^2 / (\bar{C}(\omega)^2 + \bar{N}(\omega)^2), \quad (4.14)$$

where, $\bar{C}(\omega)$ is averaged in time domain and transformed in frequency domain calibration signal, $\bar{N}(\omega)$ is powered wise averaged in frequency domain noise signal. The 350 μs LFM signal pulse compression using inverse filter is in Fig. 75.

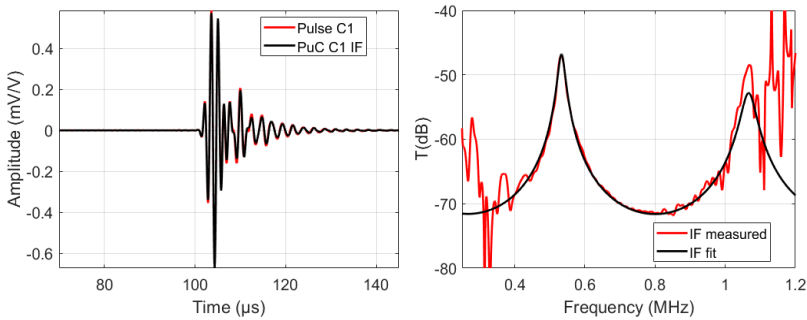


Fig. 75. (left) calibration signal using 350. 350 μs LFM signal with inverse filter using pulse compression, (right) corresponding sample transmission

The good thing about using the inverse filter is that there are no sidelobes in front which are similar to results using pulse excitation signal, while frequency response is much better with increased frequency range beyond 0.8 MHz where transduction sensitivity is low. Obtained signals correspond to excitation containing uniform spectrum, i.e. wideband Dirac pulse (actually-unity pulse).

The problem with such is that spectral programming is no longer possible, excitation signal spectra is shaped if compensation of the spectral losses, or stronger contribution of some frequency range is required. However, the use of inverse filter results in calibration signal conversion to system AC response, as shown in Fig. 76.

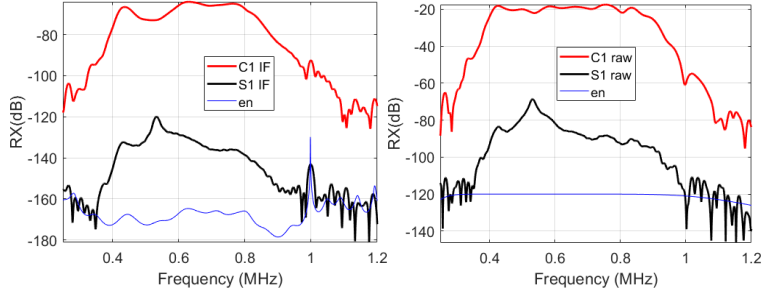


Fig. 76. (left) calibration signal using 350. 350 μ s LFM signal with inverse filter using pulse compression, (right) corresponding sample transmission

While raw signals have flat spectra as intended, the pulse compression using inverse filter destroys the flat spectra. However, SNR remains same and since fitting is done in time-domain, more focus will be provided on the transmission peak frequencies.

4.3.1.3. Pulse decompression

Gating calibration and sample is possible if sufficient damping is achieved at the overlapping regions, as calibration signal is short, the overlapping reflections will be twice further. Sample signal is sufficiently damped at 180 μ s, with no correlation sidelobes present that can be gated. Therefore, after gating, decompression is achieved using inverse of the used processing:

$$C_{IF}(\omega) = C_{IF}(\omega) \cdot G(\omega) \quad (4.15)$$

$$S_{IF}(\omega) = S_{IF}(\omega) \cdot G(\omega). \quad (4.16)$$

Filter used in Eq. 4.10, 4.11, is no longer needed due to noise gain.

4.3.2. Performance evaluation

Three different sets of signals were used for the performance evaluation: i) LFM (250 kHz-1.1 MHz), ii) APWP (380 kHz-900 kHz) optimized for flat spectrum at -20 dB, iii) APWP (380 kHz-900 kHz) to match the transducer AC response within -

20 dB. For each signal, several durations were used: 10 μ s, 20 μ s, 50 μ s, 100 μ s, 200 μ s and 350 μ s. Excitation voltage for calibration signals were 10 V, while for sample 10 V, 20 V and 50 V were used. Received signals were digitally amplified using programmable gain amplifier within 0.1-3 MHz. Calibration used 16 dB gain while sample used 45 dB (200 V excitation) to 62 dB (10 V excitation). Signals were collected using 10-bit 12.5 MHz sampling frequency ADC.

Simulations were carried out using 18-th order IIR filter, derived from LFM signal for calibration. Sample signal, using $h = 2.05$ mm, $c_s = 2240$ m/s used, $n = 1$, $\rho_s = 1193.6$ kg/m³ and $\alpha_0 = 43.5$ Np/m @650 kHz). Noise was coloured, using previously measured noise spectral density at system input.

Ten different pulse compression types were investigated: i) raw signal (labelled: Raw), ii) pulse compression using inverse filter (labelled: IF), iii) pulse compression using matched filter (labelled: MF), iv) pulse compression using mismatched filter (labelled: MMF), v) pulse compression using Wiener filter (labelled: WF), vi) raw signal with overlapping reflections reduced (ungated), vii) pulse compression using inverse filter and decompression (labelled: IFd), viii) pulse compression using matched filter and decompression (labelled: MFd), ix) pulse compression using mismatched filter and decompression (labelled: MMFd), x) pulse compression using Wiener filter and decompression (labelled: WFd). In the vi) case, compressed inverse filter signals were first gated and then decompressed, thus subtracting a raw signal. This way overlapping reflections were removed; therefore, no gating was required for decompression.

For both experimental and simulation setups, 100 measurements were measured with each excitation signals, these were then processed with each of ten compression methods and lastly using inverse solution NC-RUS fitting, sample parameters were estimated to calculate random and bias errors.

4.3.2.1. Simulation results

Pulse compression advantages can be seen in Figures 77-79. Hilbert envelope of compressed signal was taken, to make clear presentation logarithmic scale was selected. ‘Ref’ signal labelled in the results represents unity pulse.

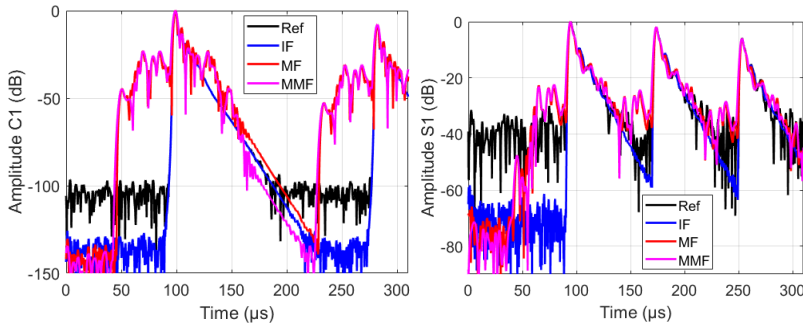


Fig. 77. Hilbert envelope of (left) calibration and (right) sample signals in case of 50 μ s duration LFM excitation

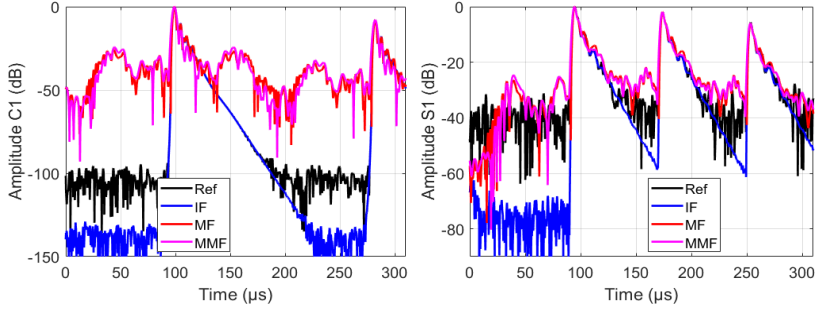


Fig. 78. Hilbert envelope of (left) calibration and (right) sample signals in case of 100 μs duration LFM excitation

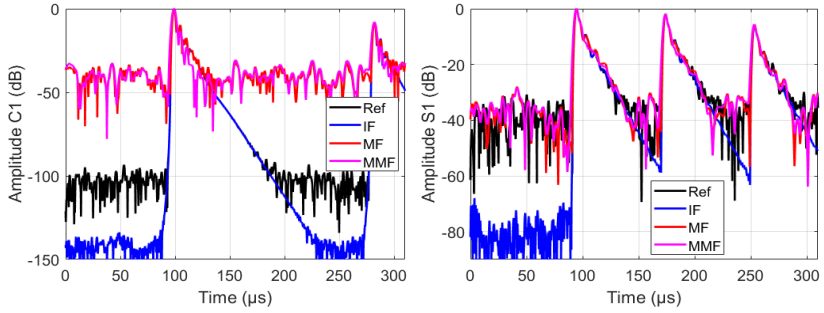


Fig. 79. Hilbert envelope of (left) calibration and (right) sample signals in case of 350 μs duration LFM excitation

The matched filter and mismatched filter in case of 50 μs , the sidelobes are only in 50 μs region, while for 100 μs and 350 μs , the sidelobes cover all ringdown tail of the signal, making proper gating or NC-RUS quite complex. The SNR is already good for 50 μs case, large SNR improvement is mainly due to large difference in the duration of signal. The difference between the lowest (10 μs 10 V LFM) and the highest (350 μs 50 V LFM) SNR is presented in Fig. 80, for the sample transmission response.

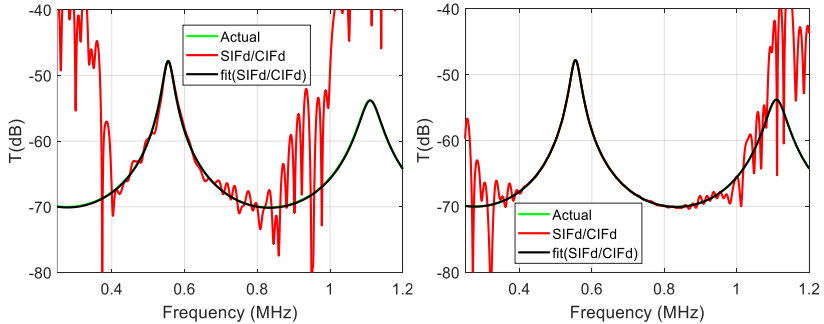


Fig. 80. Transmission response of (left) sample 10 μs 10 V LFM, (right) 350 μs 50 V LFM

The 10 μ s case shows reasonable SNR only around the peak area, while 350 ms provide good SNR even around valley area (850 kHz). However, the experimental results, 50 V excitation can produce nonlinearities in air and fitting performance at low SNR is similar for all pulse compression cases (Fig. 81).

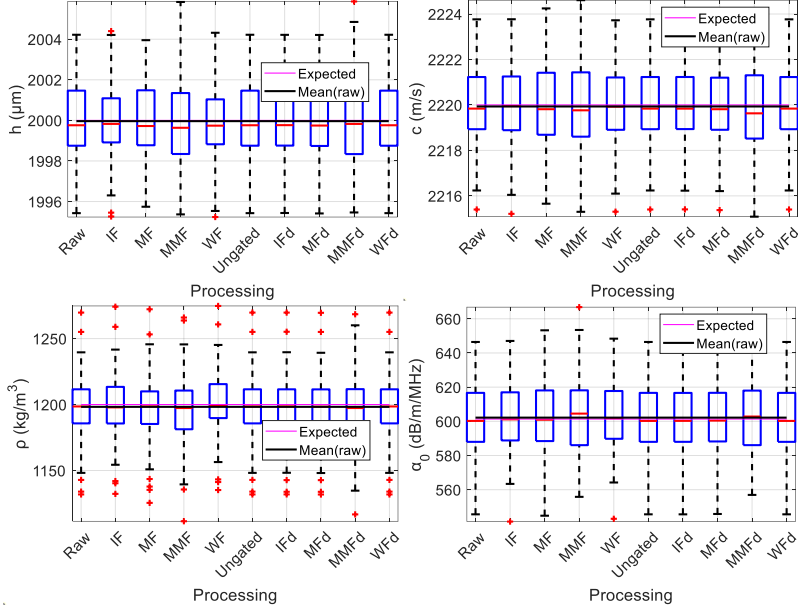


Fig. 81. Parameter excitation of NC-RUS for 10 μ s 10 V LFM

It can be clearly observed the bias errors are highest for matched filter, mismatched filter, and mismatched filter decomposition cases. The inverse filter, raw and ungated cases are similar to each other as LFM signal duration is short and there is no overlapping. Bias errors are presented in Fig. 82, it can be seen that mean is not estimated reliably: expected variation (black whiskers) of mean (with number of measurements accounted) is larger than bias error itself.

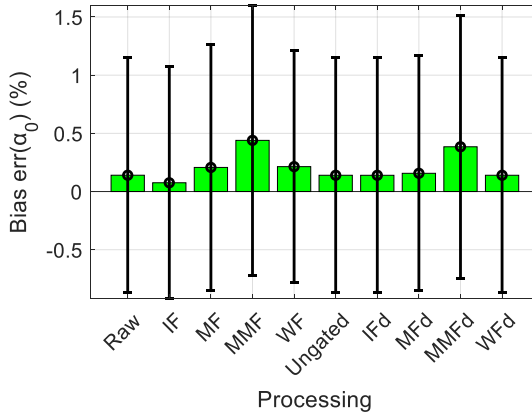


Fig. 82. LFM10 μ s attenuation α bias

For the long duration $100\ \mu\text{s}$ and $10\ \text{V}$ (Fig. 83), the results are already better despite an overlap of sidelobes, thus, the performance of Raw, MF, MMF and MMFd suffers.

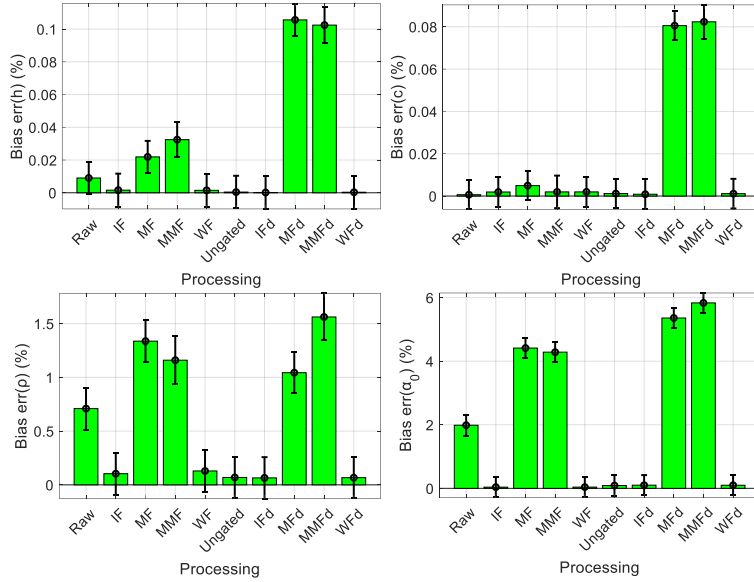


Fig. 83. Bias errors for $100\ \mu\text{s}$ $10\ \text{V}$ LFM

For the longest duration, $350\ \mu\text{s}$ the overlap of sidelobes is much more and bias errors for Raw, MF, MMF and MMFd are worse compared to IF, WF, IFd and WFd which are quite low at same level (Fig. 84).

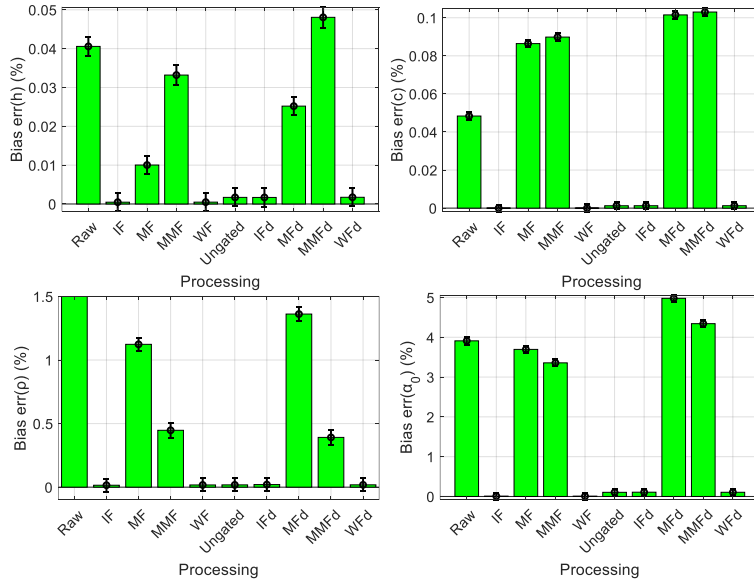


Fig. 84. Bias errors for $350\ \mu\text{s}$ $10\ \text{V}$ LFM

It is already proven that long duration sequence, especially at high voltage level contains extra SNR margin in passband, which allows the spectral programming to further improve the SNR in regions where it is low. This can be achieved by using APWP signals with spectral losses compensated (Fig. 85 left and Fig. 86 left). Therefore, signal spectra can be programmed to match the system response by concentrating energy to regions with good sensitivity (Fig. 85 right and Fig. 86 right).

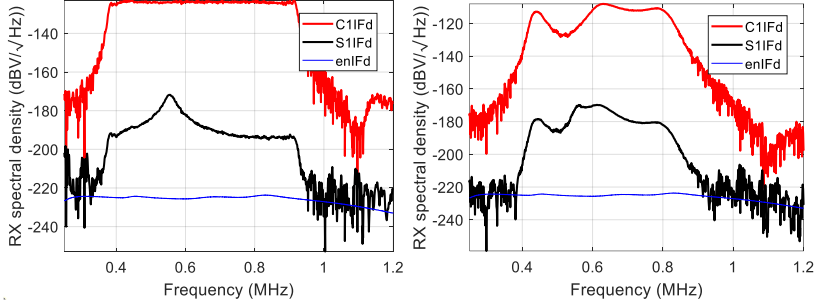


Fig. 85. (left) Spectral losses compensated spectra of calibration and sample with 50 V APWP of 350 μ s duration, (right) transmission match with 50 V APWP of 350 μ s duration

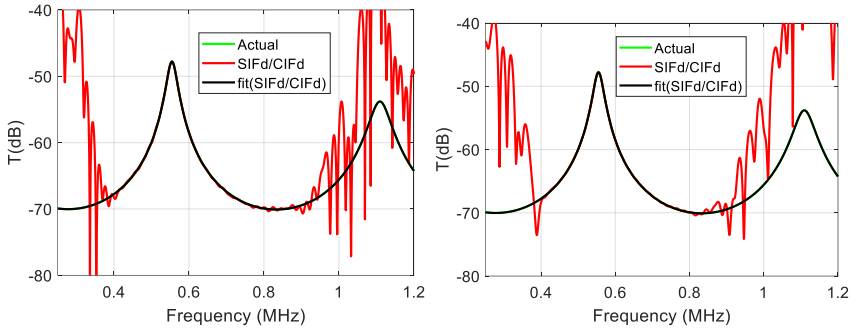
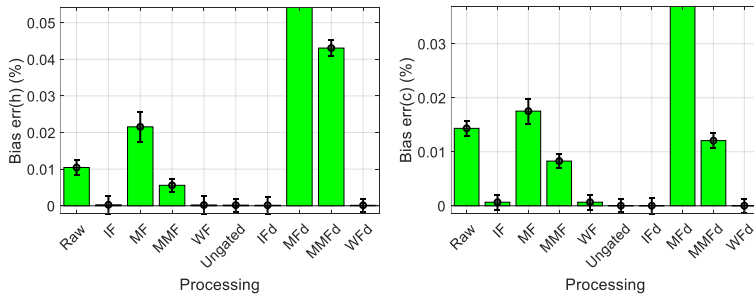


Fig. 86. (left) Spectral losses compensated for AC response of sample transmission with 50 V APWP of 350 μ s duration, (right) transmission match with 50 V APWP of 350 μ s duration

The bias errors for NC-RUS using spectral losses compensated for 50 V APWP of 350 ms duration are shown in Fig. 87.



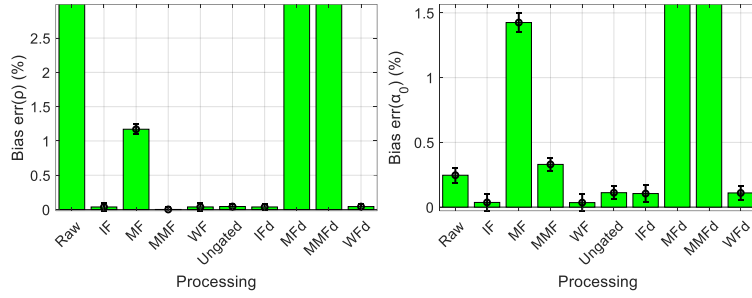


Fig. 87. Bias errors for 350 μ s 50 V APWP

The bias errors for NC-RUS using spectral losses compensated and transmission match for 50 V APWP of 350 μ s duration is shown in Fig. 88.

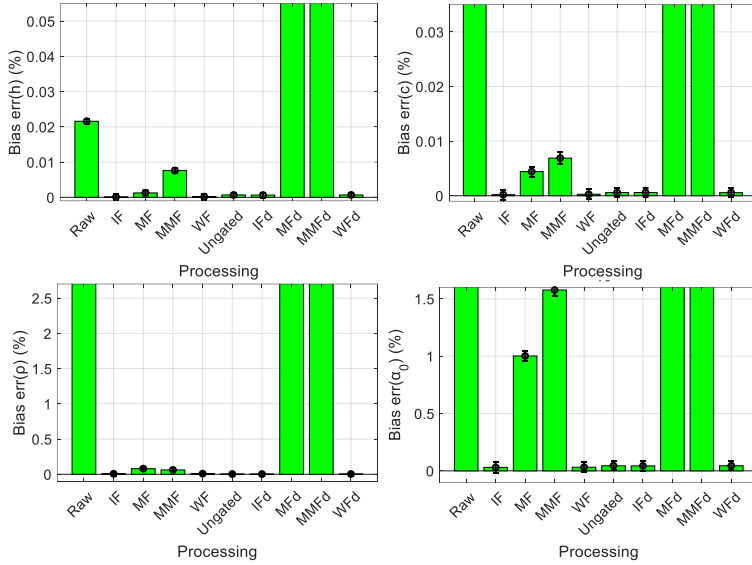


Fig. 88. Bias errors for transmission match 350 μ s 50 V APWP

In case of transmission-matched spectrum performance of Raw, MF, MMF, MFd and MMFd is low: bias error for density for MFd is 182%, 25% for MMFd, attenuation is estimated with 43% bias for MFd, 3% for MMFd. It can be noted that thickness and velocity estimation errors are very low. The reason is that these parameters are mainly estimated by delay, and delay-based measurements always have lower errors.

It can be concluded that both compressed and decompressed IF and WF pulse compression perform the best. Ungated measurement is equal to IF performance, therefore can be used as ground truth in real experiments.

4.3.2.2. Experimental results

The experimental measurements are quite complex and include many environmental and equipment limitations; therefore, several measures were taken to reduce the bias errors. Varying temperature (although in a laboratory-controlled environment not much change) was compensated using temperature compensation method [43], special ADC were used for voltage measurements, and preamplifier gain was calibrated. Due to small bias errors, it was decided to use ungated measurement results as a reference for the sample parameter estimation. Bias errors for 20 V, 10 μ s are presented in Fig. 89

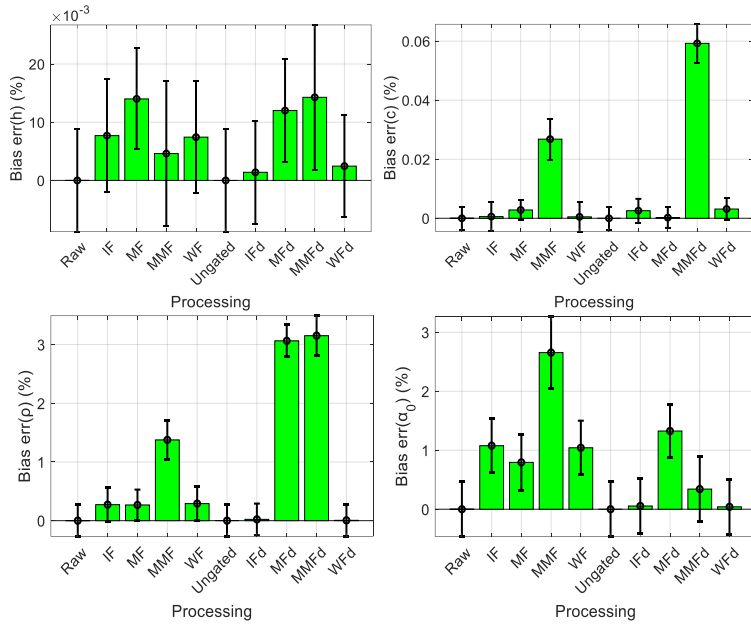
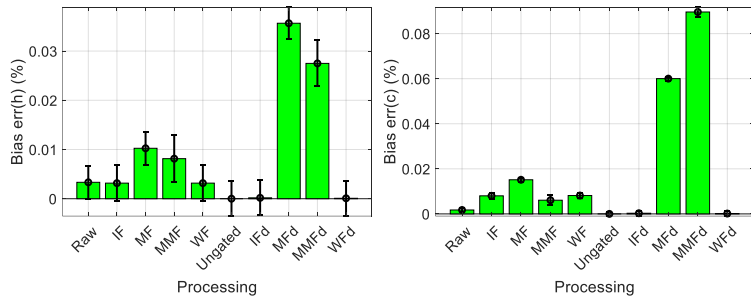


Fig. 89. Bias errors for 10 μ s 20 V LFM

The bias errors for the 20 V, 100 μ s duration are presented in Fig. 90



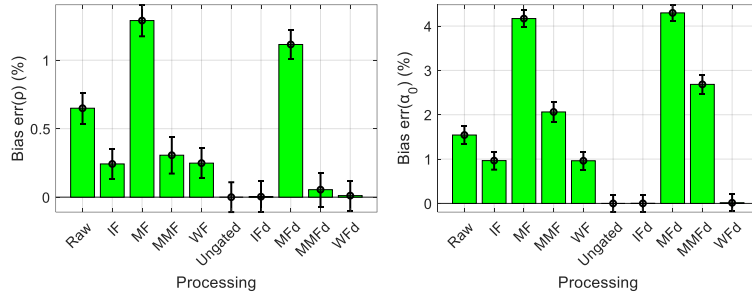


Fig. 90. Bias errors for 100 μ s 20 V LFM

The bias errors for the 20 V, 350 μ s duration are presented in Fig. 91

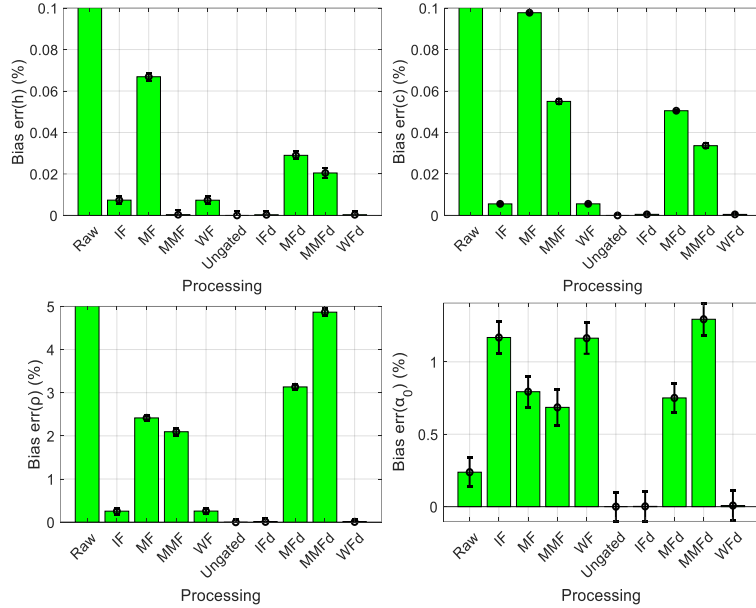


Fig. 91. Bias errors for 350 μ s 20 V LFM

Through experimental investigations, it is observed the bias errors for MF, MMF, and MMFd based pulse compression increased with longer duration signals.

4.4. Conclusions

The demand of long duration spread spectrum signals is increasing in NC-RUS due to non-linearities in air and impedance mismatch, SS signals expand the bandwidth that is available for analysis and achieve high SNR. Due to limited distance between transducers and long duration of SS signals, reflections in ultrasound measurements overlaps, which makes it difficult to use the signals for processing in inverse solution.

Two different approaches were used for solving the overlapping signals reflections: deep neural network based overlapping reflections and inverse filter-based pulse compression for overlapping reflections.

Res-UNet1D based deep learning approach was proposed for the long-duration reflection overlapping in chirp signals used for air-coupled ultrasound. Training was done using short duration signals and testing was done for both short and long duration signals. The deep learning approach was only tested for calibration signal modelling; however, the novelty is that such approach has never been tried before in literature, future studies will focus on transmission response or sample measurements for NC-RUS.

Inverse filter-based pulse compression method was proposed and tested for NC-RUS using simulations and experimental analysis. Bias errors can be reduced significantly using inverse filter with compressed signal gating and decompression.

5. PERFORMANCE EVALUATION IN THICKNESS RESONANCE ULTRASOUND SPECTROSCOPY

Now, with complete method for signal optimization and processing available, performance of thickness resonant ultrasound spectroscopy (TRUS) can be evaluated. Processing will use long, overlapping spread spectrum signals. Signals will be compressed using inverse filter, gated then decompressed back and submitted for TRUS processing as it was described in section IV. The 2 mm thick polycarbonate (PC) sample was used in investigation. First thickness resonance for such sample is expected at 560 kHz. The digital micrometre screw gauge (resolution of 0.001 mm) measured sample thickness was 2045 μm , with an expected velocity of 2240 m/s and a density of 1193.6 kg/m³ (determined from geometric and mass measurements). The reference attenuation coefficient α_0 at 560 kHz was estimated at 43.5 Np/m with a power-law exponent $n \approx 1$. Two distinct sets of experiments were carried out: i) NC-RUS, where sample is studied using air-coupled ultrasound and ii) immersion TRUS, where sample was immersed in water.

5.1. Transducers used

Air-coupled setup used same pair (Fig. 92, left) of wideband 650 kHz centre frequency transducers as in experiments above (designed and manufactured by the Spanish National Research Council, CSIC) [141]. Immersion setup used a pair of wideband composite 2 MHz centre frequency transducers (Olympus C542-SM). Transducers have 6 mm active element.



Fig. 92. Transducers used in investigation: (left) air-coupled and (right) immersion

Through transmission spectral response (derived from Calibration measurement as ratio of received to transmitted voltage) of these transducers is in Fig. 93.

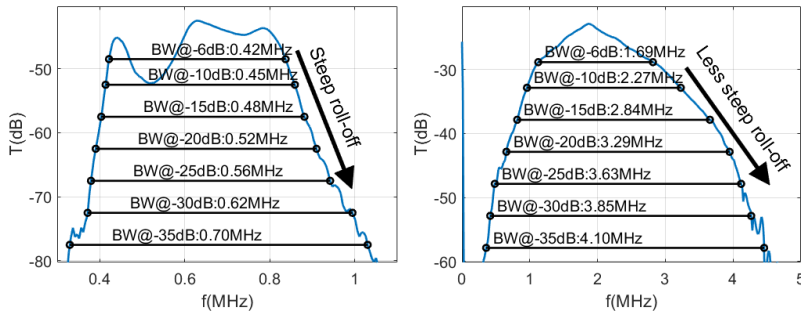


Fig. 93. AC transmission response: (left) air-coupled and (right) immersion

It can be noted that peak sensitivity for immersion transducers is by 20 dB higher (-22.9 dB vs. -42.5 dB). Essential difference is in transition band. Air-coupled transmission bandwidth is 70% @-6 dB (ratio of maximum to minimum passband frequencies is $f_{\max}/f_{\min} = 2$), 88% @-20 dB ($f_{\max}/f_{\min} = 2.4$) and 121% @-35 dB ($f_{\max}/f_{\min} = 3.2$). Immersion transducers pair transmission bandwidth is 76% @-6 dB ($f_{\max}/f_{\min} = 2.1$), 168% @-20 dB ($f_{\max}/f_{\min} = 4.6$) and 271% @-35 dB ($f_{\max}/f_{\min} = 9.2$). The difference can be distinctly seen in spectral losses compensation capability: much broader bandwidth after compensation can be obtained for immersion transducers, almost three times more than that of air-coupled transducers. With higher sensitivity and similar noise spectral density, there is 20 dB more headroom in SNR for immersion transducers which can be used for compensation.

5.2. Experiment setup

In case of air-coupled setup (Fig. 94, left) transducers were placed at 20 mm distance.

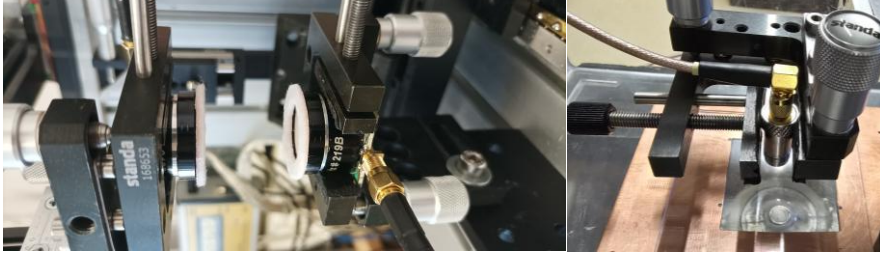


Fig. 94. Spectroscopy measurement setup: (left) air-coupled and (right) immersion

Excitation used a half bridge topology pulser (SE-TX01-02) [142], +/-10 V bipolar, rectangular chirps (linear frequency modulation, LFM) and APWP [49] sequences. Sensitivity of these transducers is very high, up to 12 Pa/V in transmission, therefore, in order to avoid nonlinearity in air and exaggerate noise effects, low excitation voltage was used. Signal from the receiving transducer was amplified by a programmable gain preamplifier SE-RX02-00 (0.1-3 MHz bandwidth, 5 k Ω input impedance) [143]. A dedicated ultrasonic signal acquisition system [143] was used both to digitize the amplifier output (10-bit analogue-to-digit converter (ADC), 100 MHz sampling rate), and to drive the pulser (binary code sets at 100 MHz sampling rate). The acquired signals were transferred to PC.

The APWP signals were derived with aim to achieve flat spectrum of the signal received in calibration measurement. For this purpose, only calibration setup was used and broadband, 100 kHz to 1.5 MHz 100 μ s long LFM excitation. The acquired signal was used to derive the 100 μ s long bipolar APWP signals using technique described in [43]. Optimization of APWP signals aimed to get a flat spectrum within predefined frequency range. Tukey (tapered cosine) window with 0.1 transition rate was used with flat portion of the window corresponding to the desired frequency range. Six signals were prepared for investigation: i) wideband 100 μ s long LFM, covering 0.3-

1.1 MHz bandwidth (labelled WB in result plots); ii) 100 μ s long LFM, covering 0.38-0.9 MHz which corresponds to -20 dB transducers bandwidth (refer Fig. 93, left, label ch20dB); iii) 100 μ s long APWP with losses compensated for 0.38-0.9 MHz bandwidth (same as ii, label AW20dB); iv) 100 μ s long LFM, covering 0.41-0.81 MHz which corresponds to -6 dB transducers bandwidth (label ch6dB); v) 100 μ s long APWP with losses compensated for 0.41-0.81 MHz bandwidth (same as iv, label AW6dB); vi) wideband, 500 ns pulse (label Pu500ns).

These signals were used to collect 100 Calibration and Sample signals which were later used in used NC-RUS measurements to estimate the PC sample parameters.

In case of immersion setup (Fig. 94, right) transducers were placed at 10 mm distance. Near field condition is satisfied only for frequencies beyond 1.5 MHz, though frequencies down to 300 kHz were used in TRUS. No diffraction correction was used. Excitation used a half bridge topology pulser (SE-TX01-02) [142], ± 0.6 V bipolar, rectangular chirps (linear frequency modulation, LFM) and APWP [49] sequences. Signal from receiving transducer was amplified by a programmable gain preamplifier SE-RX02-00 (0.1-3 MHz bandwidth, 5 k Ω input impedance) [143]. A dedicated ultrasonic signal acquisition system [143] was used both to digitize the amplifier output (10 bit analogue-to-digit converter (ADC), 100 MHz sampling rate), and to drive the pulser (binary code sets at 100 MHz sampling rate). The acquired signals were transferred to host computer.

The APWP signals were derived with aim to achieve flat spectrum of the signal received in calibration measurement. For this purpose, only calibration setup was used and broadband, 300 kHz to 3.9 MHz 40 μ s long LFM excitation. The acquired signal was used to derive the 100 μ s long bipolar APWP signals using technique described in [36,42]. Optimization of APWP signals aimed to get a flat spectrum within the predefined frequency range. Six signals were prepared for investigation: i) wideband 40 μ s long LFM, covering 0.3-3.9 MHz bandwidth (labelled WB in result plots); ii) 40 μ s long LFM, covering 0.74-3.43 MHz which corresponds to -20 dB transducers bandwidth (refer Fig. 93, left, label ch20dB); iii) 40 μ s long APWP with losses compensated for 0.74-3.43 MHz bandwidth (same as ii, label AW20dB); iv) 40 μ s long LFM, covering 1.12-2.34 MHz which corresponds to -6 dB transducers bandwidth (label ch6dB); v) 40 μ s long APWP with losses compensated for 1.12-2.34 MHz bandwidth (same as iv, label AW6dB); vi) wideband, 100 ns pulse (label Pu100ns).

These signals were used to collect 100 Calibration and Sample signals which were later used in used NC-RUS measurements to estimate the PC sample parameters and corresponding errors statistics.

5.3. Results of air-coupled NC-RUS measurements

Spectra of excitation signals used in air-coupled NC-RUS measurements is presented in Fig. 95.

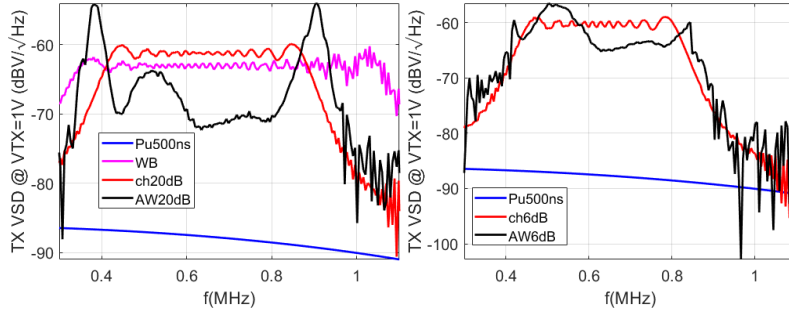


Fig. 95. Excitation signals used in air-coupled NC-RUS measurements spectra for (left) wideband and (right) narrowband set

Essence of compensation concept can be seen here: excitation signals contain more energy at transition regions (refer Fig. 95, left). Note that plot contains spectra normalized to excitation voltage (equivalent to spectra received after 1 V excitation).

Received Calibration and Sample signals spectra are presented in Fig. 96.

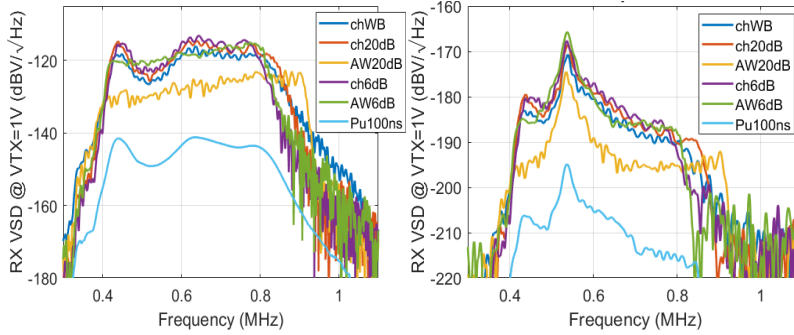


Fig. 96. Received (left) calibration and (right) sample signal spectra

It can be noted that calibration signal for LFM and pulse excitation is nonuniform, but APWP optimization has made spectrum flat [139]. APWP signal SNR is lower for the range where LFM SNR is high, but higher than LFM at frequencies where spectral losses occur (0.4 MHz and 0.9 MHz). Meanwhile, Sample signals contain the imprint of sample resonance response. Note, that APWP 20 dB spectrum (so the high SNR range available for inverse solution) is broader.

Sample transmission response at 50 V excitation for Sample, obtained after spread spectrum signal compression, using inverse filter, gating and decompression is presented in Fig. 97 to Fig. 101. Black curves correspond to maximum spectrum deviation over 100 measurements; magenta curve is the inverse solution fitting result.

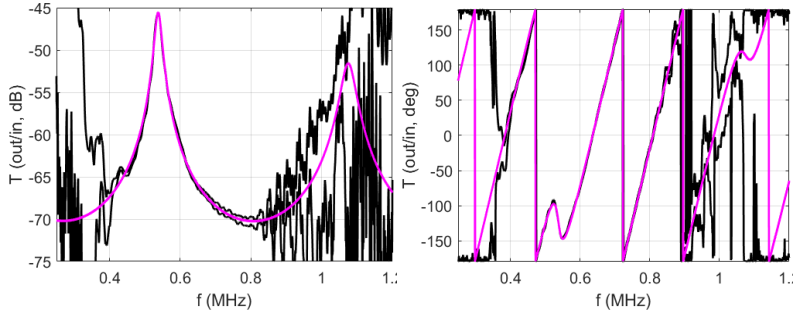


Fig. 97. Sample transmission response (left) magnitude and (right) phase for wideband LFM (label WB) at 50 V excitation

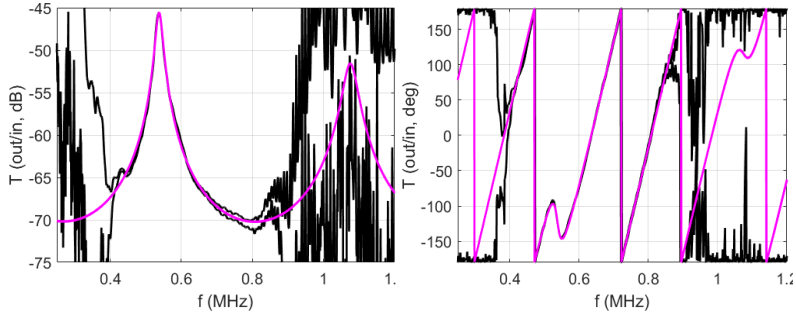


Fig. 98. Sample transmission response (left) magnitude and (right) phase for LFM bandwidth matched to -20 dB passband (label ch20dB) at 50 V excitation

It can be noted that significant spectral distortions occur beyond 0.8 MHz in both figures due to low SNR.

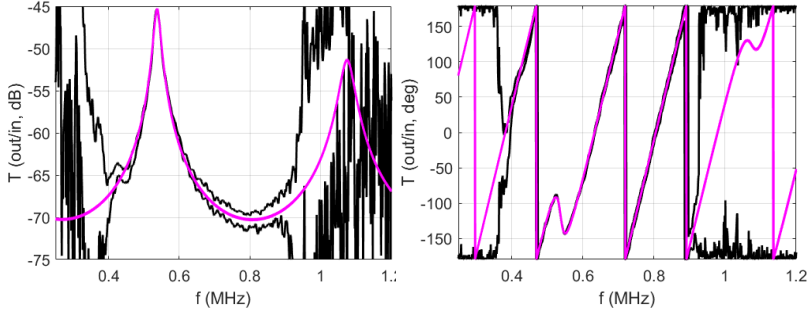


Fig. 99. Sample transmission (left) magnitude and (right) phase for APWP with losses compensation at -20 dB passband (label AW20dB) at 50 V excitation

It can be noted that spectral response is much smoother beyond 0.8 MHz in exchange to larger variation at valley region.

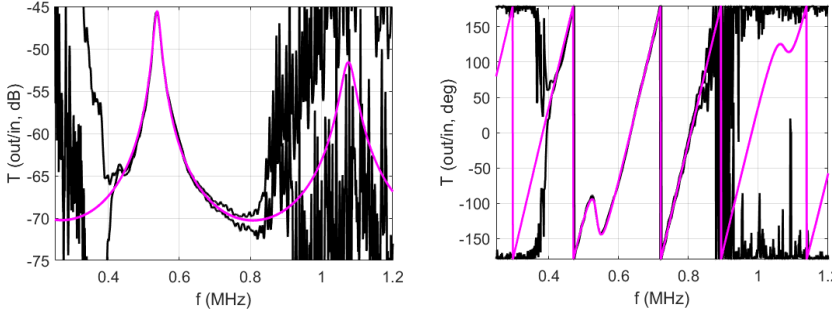


Fig. 100. Sample transmission response (left) magnitude and (right) phase for LFM bandwidth matched to -6 dB passband (label ch6dB) at 50 V excitation

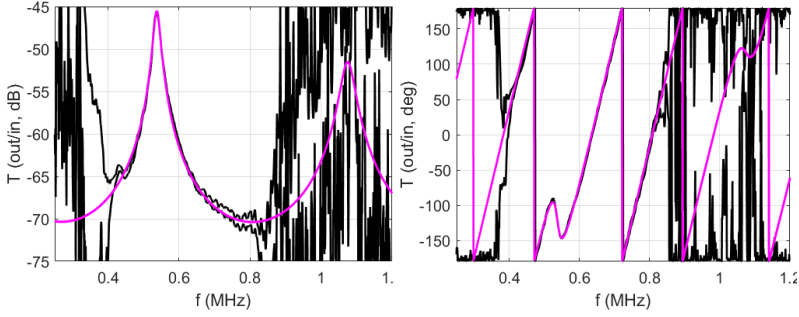


Fig. 101. Sample response (left) magnitude and (right) phase for APWP with losses compensation within -20 dB passband (label AW6dB) at 50 V excitation

It can be noted that narrowband versions cannot provide spectral information beyond 0.8 MHz. This is natural to expect, since bandwidth covered is 0.41-0.81 MHz (matched to -6 dB passband).

Sample transmission response for wideband excitation at 100 V excitation for Sample, obtained after spread spectrum signal compression, using inverse filter, gating and decompression is presented in Fig. 102 to Fig. 105.

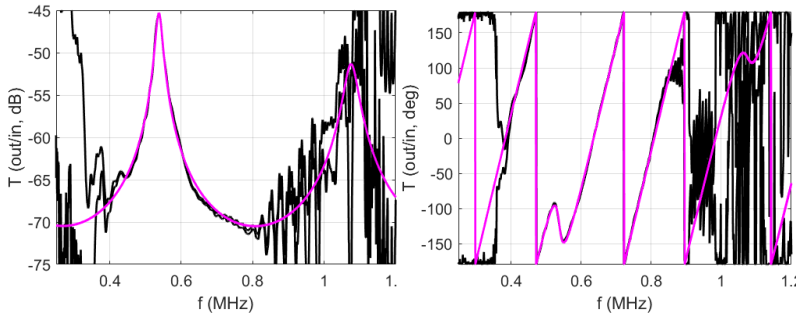


Fig. 102. Sample transmission response (left) magnitude and (right) phase for wideband LFM (label WB) at 100 V excitation

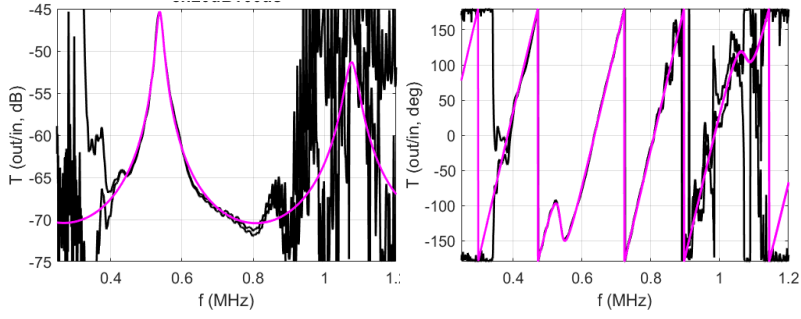


Fig. 103. Sample transmission response (left) magnitude and (right) phase for LFM bandwidth matched to -20 dB passband (label ch20dB) at 100 V excitation

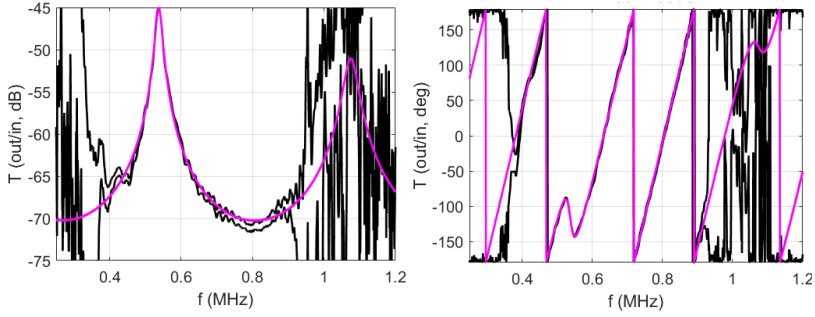


Fig. 104. Sample transmission (left) magnitude and (right) phase for APWP with losses compensation at -20 dB passband (label AW20dB) at 100 V excitation

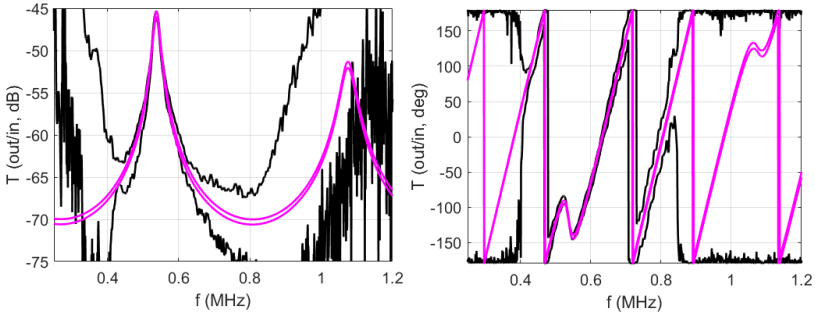


Fig. 105. Sample transmission response (left) magnitude and (right) phase for 500 ns pulse (label Pu500ns) at 100 V excitation

Problem of spectral distortion beyond 0.8 MHz remains for LFM signals, while improved behaviour remains for APWP but with reduced variation in valley region. Results obtained with pulse excitation are included for comparison. Significant difference in SNR performance for pulse excitation can be noted.

Sample magnitude transmission response for wideband LFM and APWP at 20 V excitation is presented in Fig. 106.

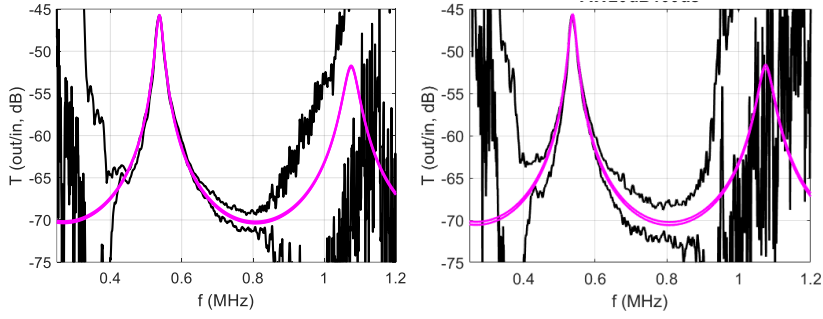


Fig. 106. (left) Sample magnitude transmission response for wideband LFM (label WB) and (right) for APWP with losses compensation at -20 dB passband (label AW20dB) at 20 V excitation

It can be noted that spectral variations are large at valley region, but spectral response is smoother beyond 0.8 MHz for APWP excitation.

Resulting NC-RUS inverse solution errors at 100 V excitation are presented as whisker plots in Fig. 107.

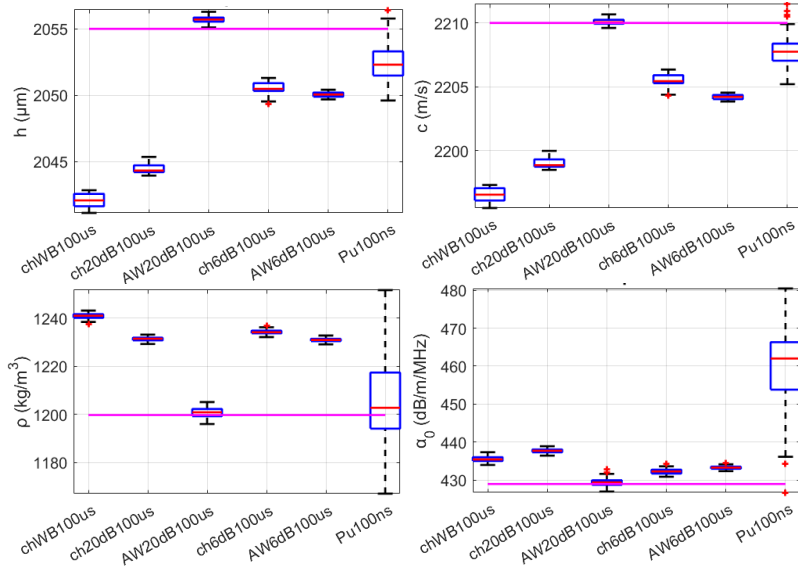


Fig. 107. NC-RUS errors for (top left) sample thickness h , (top right) velocity c , (bottom left) density ρ and (bottom right) attenuation α at 100 V excitation

It can be noted that lowest bias errors are obtained for excitation using wideband APWP with losses compensation at -20 dB passband (label AW20dB). Random errors are the lowest for narrowband APWP with losses compensation at -6 dB passband (label AW6dB). Situation is slightly different for 20 V (lower SNR) excitation, see Fig. 108.

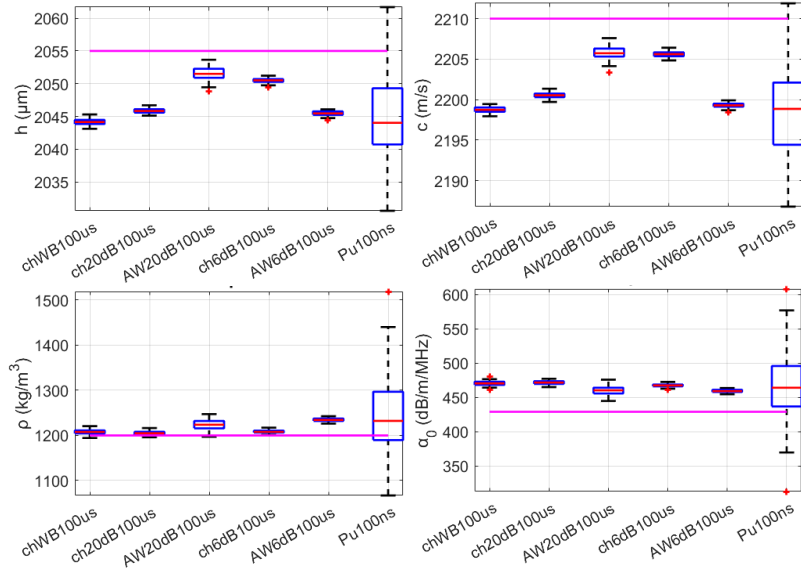


Fig. 108. NC-RUS errors for sample (top left) sample thickness h , (top right) velocity c , (bottom left) density ρ and (bottom right) attenuation α at 20 V excitation

It can be noted that bias errors are the lowest for wideband APWP with losses compensation at -20 dB passband (label AW20dB) only for velocity and thickness. Meanwhile, errors for density and attenuation are worse for this signal. The reason is that parameters that depend on amplitude (density and attenuation) at insufficient SNR suffer from increased errors. In turn, parameters that are estimate form time measurements (velocity and thickness) are more immune for SNR reduction.

Separated bias and random errors for thickness are in Fig. 109 and Fig. 110.

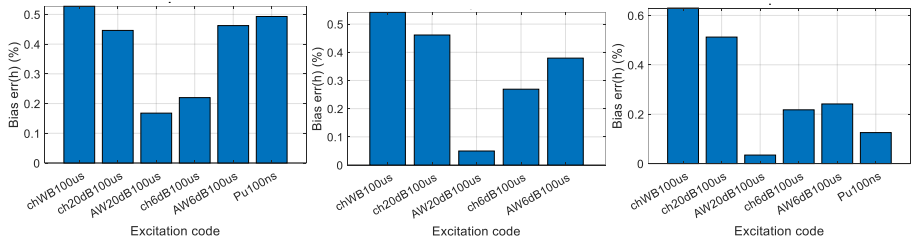


Fig. 109. Bias NC-RUS errors for sample thickness h estimation at (left) 20 V excitation, (centre) 50 V excitation and (right) 100 V excitation

It can be noted that random errors are larger for wideband APWP with losses compensation at -20 dB passband (label AW20dB) at low SNR, yet difference form other spread spectrum signals is not large and errors are quite small, 0.05% to 0.15% range. Main contribution to measurement uncertainty will come from bias errors.

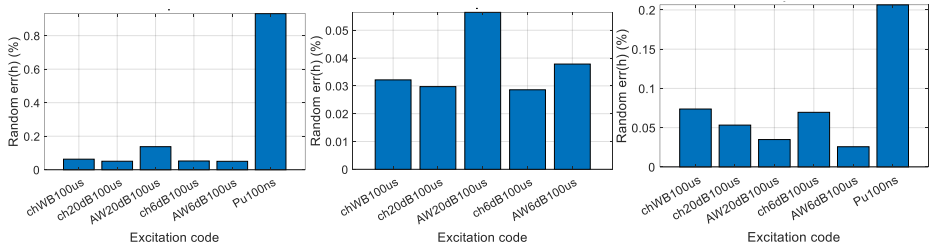


Fig. 110. Random NC-RUS errors for sample thickness h estimation at (left) 20 V excitation, (centre) 50 V excitation and (right) 100 V excitation

5.4. Results of immersion TRUS measurements

Spectra of excitation signals used in immersion TRUS measurements is presented in Fig. 111.

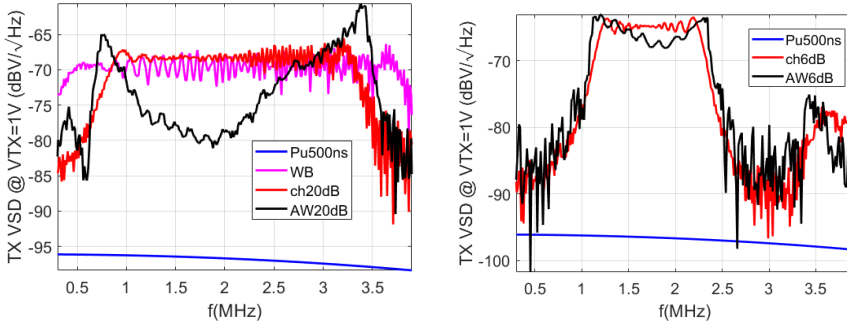


Fig. 111. Excitation signals used in immersion TRUS measurements spectra for (left) wideband and (right) narrowband set

Essence of the compensation concept can be seen here: excitation signals contain more energy at transition regions. Note that plot contains spectra normalized to excitation voltage (equivalent to spectra received after 1 V excitation). Actual excitation voltage was very low: 0.6 V. Reception gain was also small, 8 dB.

Received Calibration and Sample signals spectra are presented in Fig. 112.

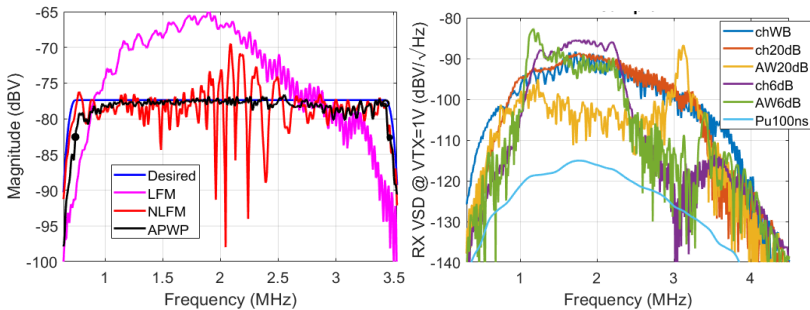


Fig. 112. Received (left) calibration and (right) sample signal spectra

NLFM signal was included in Calibration spectra plot to demonstrate the effect of APWP optimization. It can be noted that NLFM has very large spectral ripples which are smoothed out by APWP optimization. For Sample signals, it can be noted that APWP is producing higher SNR at 3.2 MHz region.

Sample transmission response, obtained after compression using inverse filter, gating and decompression is presented in Fig. 113 and Fig. 114. Black curves correspond to maximum spectrum deviation over 100 measurements; magenta curve is the inverse solution fitting result.

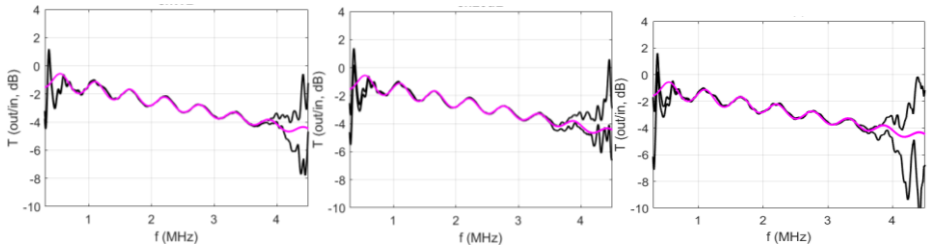


Fig. 113. Sample transmission magnitude: (left) wideband LFM, (centre) LFM -20 dB passband and (right) APWP with -20 dB passband

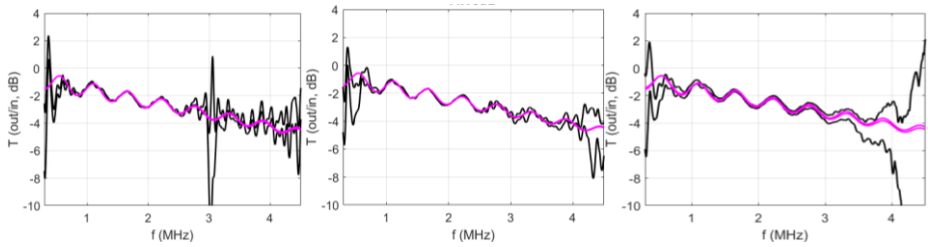


Fig. 114. Sample transmission magnitude: (left) LFM -6 dB passband, (centre) APWP -20 dB passband and (right) 100 ns pulse

Notably, significant spectral distortions occur where excitation spectral content is missing (beyond 3.5 MHz for wideband signal versions and beyond 2.5 MHz for narrowband versions). Pulse SNR degrades beyond 2 MHz. Result is expected because wideband versions cover 0.3-3.9 MHz (WB) or 0.74-3.43 MHz (ch20dB and AW20dB) and narrowband versions cover 1.12-2.34 MHz range.

Resulting TRUS errors are presented as whisker plots in Fig. 115. It can be noted that the lowest bias errors are obtained for excitation using wideband APWP with losses compensation at -20 dB passband (label AW20dB). Errors are larger for sample velocity estimation but reason for larger errors is that temperature compensation was not used in immersion measurements or polycarbonate has soaked in the water (disadvantage of immersion measurements).

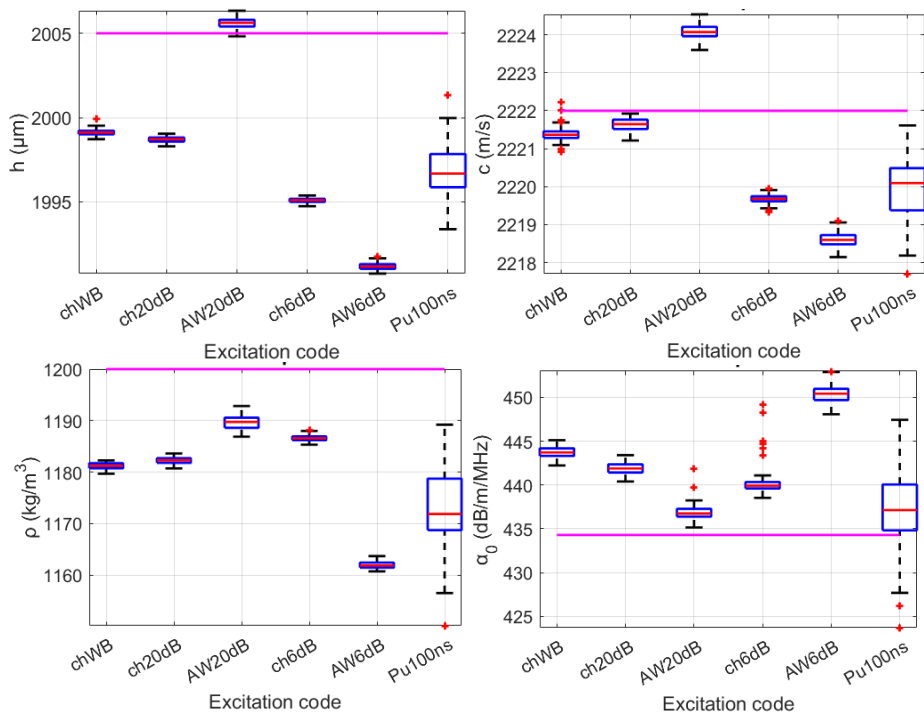


Fig. 115. TRUS errors for sample (top left) thickness h , (top right) velocity c , (bottom left) density ρ and (bottom right) attenuation α in immersion measurements

Separated bias and random errors for thickness estimation are presented in Fig. 116 and Fig. 117.

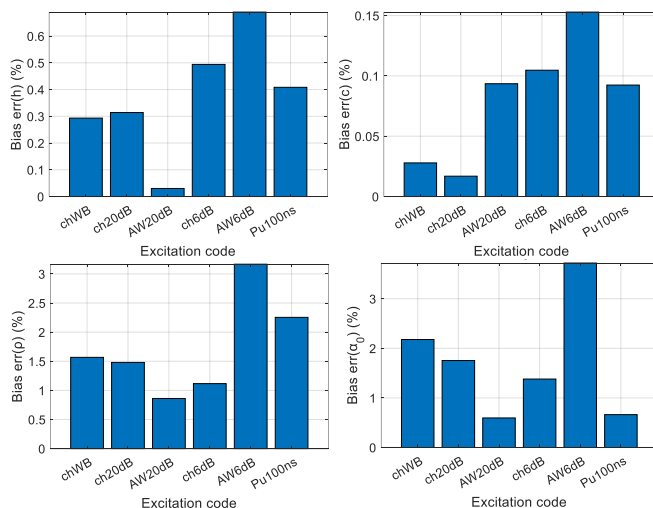


Fig. 116. Bias TRUS errors for sample (top left) thickness h , (top right) velocity c , (bottom left) density ρ and (bottom right) attenuation α

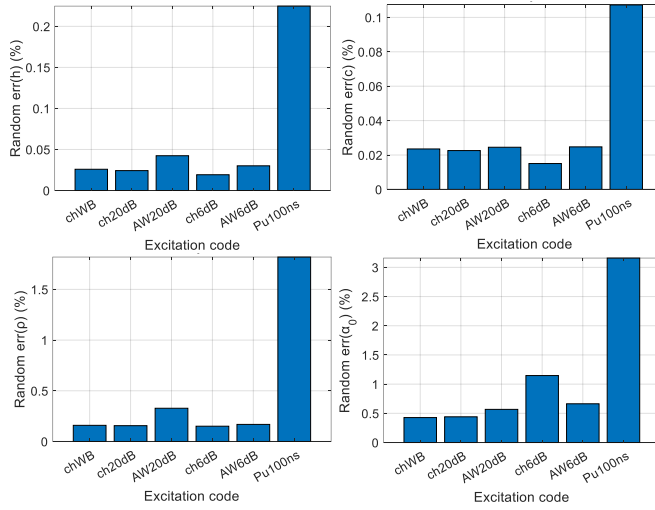


Fig. 117. Random TRUS errors for sample (top left) thickness h , (top right) velocity c , (bottom left) density ρ and (bottom right) attenuation α

It can be noted that random errors are larger for wideband APWP with losses compensation at -20 dB passband (label AW20dB) at low SNR, yet, difference from other spread spectrum signals is not large and errors for time-dependent measurements (thickness and velocity) are quite small, 0.025% to 0.04% range. Errors for amplitude-dependent measurement (density and velocity) are much larger because sample impedance is very close to water; therefore, it becomes complicated to estimate resonant peak and valley amplitude difference. Main contribution to measurement uncertainty will come from bias errors.

Narrowband excitations produced larger bias errors. Pulse excitation, as expected, showed the highest random errors due to its low energy content. Advantage of wideband, losses compensated signal can be confirmed from immersion measurements.

5.5. Conclusions

It was demonstrated that inverse filter-based pulse compression can be used in air-coupled and immersion TRUS measurements to solve long spread spectrum signals overlapping problem.

Bias errors can be reduced if APWP signal with losses compensation at -20 dB passband is used. In air-coupled measurements thickness estimation bias errors are 0.18% for wideband APWP signal (0.38-0.9 MHz) in comparison to 0.45% for same bandwidth LFM signal and 0.52% for wideband (0.3-1.1 MHz) LFM in case of 20V excitation for Sample measurement. For 100 V Sample measurement, APWP with losses compensation at -20 dB produces 0.03% bias error versus 0.45% for same bandwidth LFM signal and 0.62% for wideband LFM. Random thickness estimation errors are in 0.05% to 0.15% range for APWP with losses compensation at -20 dB

passband versus 0.03% to 0.07% range for wideband LFM signals. Larger APWP errors are expected because part of energy was used for spectral losses compensation.

In immersion measurements, thickness estimation bias errors are 0.04% for wideband APWP signal (0.74-3.43 MHz) in comparison to 0.31% for same bandwidth LFM signal and 0.3% for wideband (0.3-3.9 MHz) LFM in case of 0.8V excitation. Random thickness estimation errors are 0.04% for APWP with losses compensation at -20 dB passband versus 0.025% for wideband LFM signals.

CONCLUSIONS

1. A generalized APWP optimization framework was developed that shapes the excitation spectrum to maximize SNR within the transducer passband. Compared to conventional pulse and chirp signals, APWP increases the usable -20 dB bandwidth from ~480 kHz to 670 kHz while keeping the same peak voltage. Unipolar excitation was shown to be sufficient for spectral shaping, significantly simplifying the required electronics.

2. For TRUS applications, SNR-optimized APWP signals consistently produced the lowest bias errors for all four estimated parameters (α , c , ρ , h).

a. For amplitude-sensitive parameters (density, attenuation), APWP reduced relative bias errors by factors of 5-10 times compared to LFM and pulse excitations;

b. For time-sensitive parameters (velocity, thickness), errors were reduced by 3-15 times. The improvement is most pronounced in air-coupled conditions, where high attenuation and reverberations dominate.

3. Analysis of spectral coverage revealed a threshold effect on bias errors: when the transducer bandwidth excluded the sample resonance peak and adjacent valley, the estimated parameters showed large bias errors; once included, the bias errors decreased markedly and then remained nearly stable. Therefore, under the investigated conditions, one peak-valley pair was sufficient for stable estimation, while additional covered peaks did not result in a significant further error reduction. Coverage of only the peak leads to errors of 7-9 %. Including both valleys minimizes errors for attenuation (~3.8 %). The best overall compromise was achieved when the spectrum covers the full - 20 dB transducer passband.

4. An inverse filter-based pulse compression method was proposed and validated for decomposing overlapping reflections in long-duration spread-spectrum signals. Combined with pulse compression gating and decompression, it reduces bias errors (velocity error from 0.085 % to 0.002 % and density error from 1.2 % to 0.05 % at 350 μ s / 20 V excitation).

5. A ResidualUNet1D deep learning model was developed for automated reflection separation in long-duration spread-spectrum signals, achieving ~45 dB RE and correlation ~0.9 for 50 μ s signals, and ~32 dB RE with correlation ~0.74 for 300 μ s signals.

6. In immersion measurements, thickness estimation bias errors are 0.04% for wideband APWP signal (0.74-3.43 MHz) in comparison to 0.31% for same bandwidth LFM signal and 0.3% for wideband (0.3-3.9 MHz) LFM in case of 0.8V excitation. Random thickness estimation errors are 0.04% for APWP with losses compensation at -20 dB passband versus 0.025% for wideband LFM signals.

7. The proposed APWP spectral compensation technique, combined with inverse-filter pulse compression and deep-learning decomposition, improves the quality and accuracy of air-coupled TRUS measurements. These methods enable reliable resonance parameter extraction using only a single transducer pair across wide bandwidths, overcoming previous limitations in highly attenuative air-coupled conditions.

SANTRAUKA

IVADAS

Neardomieji bandymai (toliau – NDT) yra būtini medžiagų ir konstrukcijų saugumui, patikimumui ir ilgaamžiškumui užtikrinti daugelyje taikymo sričių [1, 2]. Iš visų NDT metodų ultragarsas išsiskiria savo gebėjimu tirti neinvaziniu būdu [3]. Tiesioginis ryšys tarp bangų sklaidimo ir medžiagos mechaninių savybių yra priežastis, dėl kurios ultragarsiniai matavimai yra tokie populiarūs [4], be to, tai yra nejonizuojanti spinduliuotė, įranga yra nešiojama, pigi, mažos galios. Ultragarsas plačiai naudojamas biomedicinoje [5], aviacijoje [6], civilinėje inžinerijoje [7], medžiagų bandymuose [8, 9], pramoniniame apdorojime [10], srauto matavimuose [11], maisto pramonėje [12, 13] ir netgi trapių ar minkštų medžiagų tyrimuose [14].

Apskritai ultragarsinis matavimas iš esmės apima elektros energijos konversiją į akustines bangas, tų bangų įvedimą į bandomąjį mėginį ir tada konversiją atgal į priimančią elektroniką [1, 2]. Atspindėtų arba perduotų signalų analizė leidžia apibūdinti medžiagą. Tokio matavimo kokybė priklauso nuo efektyvaus konvertavimo. Yra keletas ultragarso įvedimų į bandomąjį tipą: kontaktinis, imersinis, įvedimo per orą, termoakustinis, lazerinis termoakustinis, elektromagnetinis akustinis keitiklis (EMAT) [15, 16].

Tiesioginio kontakto su bandiniu atveju keitiklis ir bandinys liečiasi vienas su kitu [13]. Abiejų paviršių nelygumai lemia tai, jog akustinio kontakto kokybė neatsikartoja, o skenavimas yra sudėtingas. Todėl naudojamas kontaktinis skystis (paprastai plonas vandens, gelio arba aliejaus sluoksnis). Tarpiniu atveju, įvedant per sausą kontaktą, kontaktui užtikrinti naudojama minkšta membrana [17]. Pagrindinės kontaktinio ultragarso taikymo sritys yra defektų nustatymas, storio matavimas, suvirinimo siūlių tikrinimas, klijų tikrinimas, erozijos žemėlapių sudarymas, terminio poveikio zonos vertinimas ir aukštos temperatūros komponentų tikrinimas [2]. Kontaktinis ultragarsas užtikrina stiprų signalą dėl minimalaus akustinės varžos nesuderinamumo, o keitiklį galima greitai pritvirtinti prie bandinio. Tačiau trapūs arba šiurkštūs paviršiai gali būti netinkami tiesioginiam kontaktui, o erdvinė skiriamoji geba gali būti ribota dėl keitiklio dydžio ir spindulio plėtimosi tikrinamoje medžiagoje, mechaniniam skenavimui reikia nuolatinio kontaktinės terpės tiekimo [13]. Tačiau išlaikyti nuolatinį akustinį kontaktą yra sudėtinga: paviršiaus šiurkštumas kinta, sunku išlaikyti tinkamą keitiklio pasvirimą, paviršiaus abrazyvumas sunaikina suderinimo sluoksnio storį ir vientisumą. Keitiklių gardelės iš dalies sąlygoja minėtas problemas, tačiau apertūra (taigi ir skiriamąją gebą riboja gardelės dydis) vis dar ribojama matricos dydžio. Tačiau išlaikyti nuolatinį akustinį kontaktą yra sudėtinga: paviršiaus šiurkštumas kinta, sunku išlaikyti tinkamą keitiklio pasvirimą, paviršiaus abrazyvumas sunaikina suderinimo sluoksnio storį ir vientisumą.

Imersiniu atveju mėginys panardinamas į skystą terpę (paprastai vandenį). Panardinimas užtikrina gerą jungtį, galima naudoti fokusuotus keitikius, galima tirti sudėtingos geometrijos mėginius [18,19]. Tačiau, kadangi mėginys turi būti panardintas į skystį, tokia jungtis netinka trapioms, vandeniui jautrioms ar akytoms

medžiagoms. Be to, jai reikalinga įranga su keliais keitikliais (pvz., perdavimas per bandinį) ir sudėtingas skenavimas, o keitikliai ir įranga turi būti atsparūs vandeniui.

Įvedimas per orą naudoja orą kaip įvedimo terpę [20], todėl tai yra bekontaktis matavimas. Įvedimas per orą yra idealus porėtoms, pluoštinėms medžiagoms, įvairiems maisto produktams, biologiniams audiniams, drėgmei jautrioms ir aukštos temperatūros medžiagoms [21]. Skenavimo įranga neturi būti atspari vandeniui, galima sukonfigūruoti sudėtingas skenavimo sistemas, greitai atlikti matavimus.

Lazerinis ultragarsas naudoja impulsinį lazerį, kad įkaitintų mėginio paviršių, kuris generuoja ultragarsines bangas [22]. Lazerinė vibrometrija gali būti naudojama ultragarsinėms bangoms ant mėginio paviršiaus nustatyti [23]. Paprastai lazerio spindulio, atspindėto nuo mėginio, Doplerio poslinkis naudojamas paviršiaus poslinkio kartografavimui su didele erdvine ir laiko skiriamąja geba. Tačiau lazerio naudojimas (plazmos režimas naudoja $\geq 10^7 \text{ mW/cm}^2$, 100 milijonų kartų daugiau nei saulės spinduliuotė) yra susijęs su akių ($2,5 \text{ mW/cm}^2$) ir odos pažeidimais (100 W/cm^2), lazerio ultragarso sistemos yra didelės, brangios, reikalauja tikslaus suderinimo ir gali būti jautrios paviršiaus sąlygoms, pvz., šiurkštumui ir atspindžio koeficientui, o tai gali riboti jų naudojimą kai kuriose pramonės srityse [24]. Lazerinei vibrometrijai reikia gerai atspindinčio paviršiaus, ji yra jautri aplinkos vibracijoms ar aplinkos šviesos trukdžiams.

EMAT generuoja ir aptinka ultragarsą elektrai laidžiose medžiagose per elektromagnetinę indukciją, todėl nereikia jungties ar fizinio kontakto [25]. Šis metodas paprastai turi mažą energijos konversijos efektyvumą, palyginti su pjezoelektriniais keitikliais, dėl to gaunamas mažesnis SNR. Be to, stiprūs magnetai gali suspausti ir išplėsti pirštus ar odą [26]. Be to, EMAT tikrinimas apsiriboja elektriškai laidžiomis medžiagomis ir gali reikalauti specialių signalų apdorojimo prietaisų [27].

Šiame darbe nagrinėjamas tik imersinis [18] ir (daugiausia) oro [28] ultragarsas. Oro ultragarsas yra tinkamesnis, saugesnis eksploatuoti, įranga lengva ir lengvai valdoma [29].

Tačiau oro akustinis impedansas yra apie 410 Rayl, tai yra daug mažiau nei daugumos kietųjų medžiagų (1–30 MRayl) [21, 30]. Ultragarso sklindant iš oro į kietą medžiagą, didžioji dalis energijos atspindima ribos paviršiuje, todėl perduodamas signalas žymiai susilpnėja sklisdamas pirmyn ir atgal (oras → kietą medžiagą → oras). Perdavimo per aliuminio bandinį slopinimo nuostoliai siekia apie 80 dB [15]. Impedanso nesuderinamumas lemia itin mažos amplitudės signalus ir santykinai mažą signalo ir triukšmo santykį (SNR). Slopinimas ore yra didelis ir priklauso nuo dažnio. Esant 100 kHz dažniui, slopinimas 100 mm oro sluoksnyje yra palyginti nedidelis (apie 1–2 dB), o esant 1 MHz dažniui jis jau yra 30–40 dB, o esant 5 MHz dažniui gali viršyti 200 dB [31]. Kartu su dideliu impedanso nesuderinamumu oro ir kietojo kūno sąsajoje, kuris gali sukelti papildomą 60–90 dB perdavimo nuostolį, bendras signalo sumažėjimas per visą laiką (oras → kietasis kūnas → oras) gali būti didesnis nei 100 dB, ypač aukštesniuose dažniuose [5, 31, 32]. Todėl oro ultragarso keitikliai paprastai yra suprojektuoti žemiems dažniams (nuo dešimčių kHz iki kelių MHz) ir veikia dešimčių milimetrų atstumu, kur signalo nuostoliai yra lengviau valdomi.

Yra du pagrindiniai ultragarso matavimų veikimo režimai: atspindys ir perdavimas. Atspindžio režimu mėginys vertinamas tik iš vienos pusės registruojant bangas, atspindėtas nuo medžiagos vidinės struktūros. Atspindėtos išilginės arba *Lamb* bangos [33] signalo skrydžio trukmė (ToF) gali būti naudojama defekto gyliui arba medžiagos storiui (jei žinomas greitis) arba greičiui (iš žinomo storio) matuoti [34]. Atspindžio amplitudė gali būti naudojama impedanso nesuderinamumui įvertinti. Atspindžio režimas, išskyrus *Lamb* bangų signalus [33], retai naudojamas oro jungiamuose matavimuose, nes didžioji dalis energijos ateina iš paviršiaus, užgoždama vidinių sluoksnių signalus. Atspindžio režimas taip pat siejamas su negyvos zonos problema, jei tas pats keitiklis naudojamas perduoti ir priimti [35]. Perdavimo režimas naudoja vieną keitiklį vienoje mėginio pusėje –signalui perduoti ir kitą keitiklį priešingoje mėginio pusėje – signalui priimti [36]. Toks nustatymas išryškina signalo savybes, susijusius su mėginio vidine struktūra, todėl galima tirti medžiagos savybes. Paprasčiausiu atveju signalo amplitudė gali būti naudojama vidiniams defektams įvertinti [34]. Tačiau ToF daugkartinių atspindžių mėginyje gali būti naudojamas medžiagos storiui (jei žinomas greitis) arba greičiui (iš žinomo storio) vertinti.

Paprastai ToF matuojamas tarp dviejų skirtingų signalų. Tuomet norint išskirti du signalus pageidautina naudoti aukštesnius dažnius. Skyrą išilgai sklidimo ašies lemia absoliutinis dažnių juostos plotis, tačiau kadangi keitiklio dažnių juostos plotis yra tam tikra keitiklio centrinio dažnio dalis, aukštesnio dažnio keitiklio naudojimas užtikrina platesnį dažnių juostos plotį. Aukštesni dažniai yra svarbūs geresnei šoninei raiškai.

Rezonansinė ultragarso spektroskopija (RUS) yra alternatyvus būdas medžiagos mechaninėms savybėms nustatyti, kuomet bandinys yra plonesnis nei skyra išilgai sklidimo ašies. Ji remiasi natūraliais mėginio vibracijos rezonansais (savaisiais dažniais), kad nustatytų jo medžiagos mechanines savybes [37–40]. RUS analizuoja medžiagos rezonansinį atsaką. Standartinėje RUS konfigūracijoje sužadinimo signalas, pvz., slenkanti sinusinė banga arba plačiajuostis impulsas, perduodamas į mėginį naudojant pjezoelektrinį keitiklį arba kitą tinkamą sužadinimo signalą. Kintant įėjimo signalo dažniui, mėginys reaguoja vibruodamas savo natūraliais rezonansiniais režimais. Susidariusios mechaninės vibracijos nustatomos priimančiuoju keitikliu arba jutikliu, kuris atsaką konvertuoja atgal į elektrinį signalą [37]. Sistemingai keičiant sužadinimo dažnį per norimą diapazoną ir registruojant atitinkamą gauto signalo amplitudę ir (arba) fazę, gaunamas rezonansinis spektras. Amplitudės spektro smailės (ir fazės spektro charakteristikos) atitinka medžiagos rezonansinius dažnius, kurie yra sudėtingai susiję su jos elastingumo, geometrinėmis ir ribinėmis savybėmis. Žinant mėginio geometriją, elastingumo savybių matrica gaunama sprendžiant atvirkštinę problemą analitinėmis išraiškomis arba baigtinių elementų metodo (FEM) modeliavimu [15, 37].

Tradicinė RUS reikalauja kruopštaus mėginio montavimo, kad būtų pasiektos laisvos, neapribotos ribinės sąlygos. Paminėtina, kad storio rezonanso ultragarso spektroskopija (TRUS) naudoja oro [21, 29] arba vandens [18] terpę signalui įvesti, kurią technologiniu požiūriu daug lengviau paruošti. Paprastai bandinio (plono lakšto) parametrams įvertinti atliekami du matavimai: a) kalibravimo matavimas, kai kelias

tarp keitiklių yra be kliūčių ir b) mėginio matavimas, kai mėginys įdedamas tarp keitiklių. Plokštės storio rezonansas sukuria rezonansinius pikus perdavimo spektruose, kurie yra susiję su mėginio storiu ir mėginio ultragarso greičiu. Be to, tarp kalibravimo ir mėginio matavimų yra signalo poslinkis laike, kuris taip pat yra susijęs su mėginio storiu ir ultragarso greičiu. Tokie du nepriklausomi matavimai leidžia vienu metu įvertinti mėginio storį, tankį, ultragarso greitį ir slopinimą, pasiekiant atvirkštinės problemos sprendimą pagal mėginio modelį. TRUS buvo sukurta tam atvejui, kai dažnių juosta yra tokia siaura, kad atspindžiai medžiagos viduje persidengia ir negalima naudoti tradicinių RUS ar ultragarso metodų. Ši galimybė leidžia TRUS būti galingu tyrimo metodu mažų ir sudėtingų objektų, įskaitant augalų lapus [32, 36, 41], charakteristikoms nustatyti, kai kontaktiniai ar panardinimo metodai yra neveiksmingi. Oro įvedimo atveju TRUS matavimas yra tikrai nekontaktinis, todėl kartais jis vadinamas nekontaktine rezonansine ultragarso spektroskopija (NC-RUS) [36,42]. Išsamus TRUS technikos paaiškinimas pateikiamas 2.2 skyriuje.

Impedanso nesuderinamumas tarp jungiamosios terpės ir bandinio sukelia didelius atspindžio nuostolius sąsajoje [43]. Taigi, per bandinį perduodama energija yra mažesnė, o tai sumažina SNR ir jautrumą elektromagnetiniams trukdžiams. Tiek jungiamosios terpės (pavyzdžiui, oro ar vandens), tiek bandomojo mėginio slopinimas kinta proporcingai dažniui. Šis poveikis ypač ryškus ore, kur slopinimas smarkiai didėja su dažniu, dar labiau sumažindamas SNR ir apribodamas praktinį aukšto dažnio ultragarso naudojimą oro įvedimo atveju. Todėl aukštesni ultragarso dažniai patiria dar didesnę silpnėjimą, o tai riboja veiksmingą tikrinimui prieinamą dažnių juostą, ypač oro jungiamosiose sistemose [43]. Paprastai [21, 29] žadinimui naudojamas trumpas impulsas, kad būtų gautas plačiajuostis sužadinimas [44]. Impulso atveju egergija ir dažnių juostos plotis atvirkščiai proporcingi, todėl siekiant skleistą spektro dažnių ruožo – SRN mažėja. Impulso spektro forma (angl. *sinc*) yra fiksuota, todėl neįmanoma selektyviai formuoti spektro ar nukreipti daugiau energijos į reikiamas dažnių juostas.

Skleistą spektro (SS) signalų juostos plotis ir energija gali būti valdomi nepriklausomai, todėl galima užtikrinti gerokai aukštesnį SNR ir platesnį dažnių ruožą. Šiame darbe naudojami *chirp* (teisinė dažnio moduliacija (LFM)) [45, 46], *netiesinis chirp* (netiesinė dažnio moduliacija (NLFM)) [47,48] ir laisvai pasirenkamos trukmės bei pozicijos impulsų sekos [36, 42, 49–51]. SS signalų spektrinė aprėptis apibrėžiama moduliacijos parametrais. SS naudojami dėl jų trumpesnių impulsų, be to, SS signalų energiją galima nukreipti į dažnius, kur nuostoliai yra didesni, todėl priimamo signalo atsaką padaro platesnį [49, 50, 52].

Tačiau iššūkis yra pats SS signalas. APWP signalo spektrinės formos lankstumas priklauso nuo signalo ilgio: kuo signalas ilgesnis, tuo spektrinė forma lankstesnė ir tikslesnė. Ilga SS signalo trukmė yra svarbi, kad būtų sukaupta pakankamai energijos plačiam dažnių diapazonui pasiekti: tam reikia paskirstyti energiją toliau nuo keitiklio centrinio dažnio, kur keitimo efektyvumas mažėja. Tačiau TRUS matavimo atveju keitikliai turi būti išdėstyti palyginti arti mėginio, kad būtų užtikrintas plokščias bangos frontas (artimo lauko veikimas) ir sumažintas oro slopinimas. Mėginio ir keitiklio paviršiaus atspindžio koeficientas yra didelis, todėl

susidaro daugkartiniai atspindžiai. Visi šie veiksniai padidina pirmojo perduoto signalo ir daugkartinių atspindžių mėginys-keitiklis persiklojimą. Dėl persiklojimo sudėtinga interpretuoti matavimo signalą, sumažėja atvirkštinio sprendinio tikslumas [49].

Sprendimas atrodo trivialus: SS signalai gali būti suspaudžiami naudojant suderintą filtrą (kryžminės koreliacijos apdorojimas). Tačiau keitiklio atsako išnykimas yra ilgas dėl kelių suderintų sluoksnių. Todėl koreliacijos šoninių spindulių lygis yra aukštas. Signalai po tokio impulsų suspaudimo negali būti filtruojami. Buvo pasiūlyta spręsti problemą dviem būdais: a) naudoti tam tikrą atvirkštinio filtro išvestinę ir b) atlikti daugkartinių atspindžių dekonvoliuciją. Siūlomas atvirkštinis filtras yra išvedamas naudojant sužadavimo genkodą (generavimo kodą) arba signalą prie siunčiančiojo keitiklio spausdukų. Atvirkštiniai filtrai turi tokį patį fazinį atsaką kaip suderinti filtrai, todėl įvyksta signalo suspaudimas. Tačiau tokio filtro taikymo rezultatas nėra suderinto filtro signalas, kuris turi simetrinius šoninius sparnus, bet yra lygiavertis dėl impulsinio sužadavimo Pasiekiami geresni arba lygiaverčiai impulsų sužadavimo rezultatai. Tada signalas gali būti filtruojamas su tokiu pačiu sėkmės laipsniu kaip ir vieno impulso signalas arba atliekamas atvirkštinio sprendimo pritaikymas laiko srityje. Rezultatai buvo palyginti su suderintu, *Wiener* ir nesuderintu filtravimu.

Vienos keitiklio poros matavimas gali tapti įmanomas esant pakankamam SNR. SS signalų gebėjimas užtikrinti aukštesnį SNR, kartu su naujų, labai efektyvių oro jungiamųjų keitiklių [21, 29, 32, 43] kūrimu, leidžia laikyti NC-RUS patikimu ir veiksmingu įrankiu medžiagų savybėms vertinti.

Hipotėzė: *Rezonansinės ultragarso spektroskopijos matavimų kokybę galima pagerinti, jei ilgos išsklaidytos spektro sužadavimo bangos yra suformuotos taip, kad jų spektrinė forma kompensuotų spektrinius nuostolius.*

Tyrimo tikslas – sukurti spektrinių nuostolių kompensavimo metodą, pagrįstą programuojamos spektrinės formos signalais ir ištirti jo veikimą rezonansinės ultragarso spektroskopijos matavimuose.

Tyrimo uždaviniai

1. sukurti transdukcijos spektrinių nuostolių kompensavimo techniką ir ištirti, ar tokia kompensacija yra veiksminga gerinant TRUS matavimus naudojant vieną keitiklių porą ir nesutampantčius signalus;

2. sukurti programuojamos spektrinės formos optimizavimo kriterijus ir ištirti, ar transdukcijos nuostolių kompensavimas yra pakankamas TRUS matavimų kokybei pagerinti, ar vis dėlto būtina atsižvelgti į mėginio dažninį atsaką;

3. sukurti ir ištirti techniką, skirtą sumažinti ilgų išsklaidytų spektro signalų persidengimo poveikį TRUS matavimuose;

4. įvertinti spektrinių nuostolių kompensavimo veiksmingumą, naudojant ilgus trukmės sužadavimo signalus TRUS matavimuose.

Mokslinis naujumas

1. Ankstesniuose tyrimuose daugiausia dėmesio buvo skiriama sužadavimo signalo spektro optimizavimui, nuostolių kompensavimui programuojant spektrinę formą [49, 51, 53]. Tačiau nebuvo atsižvelgta į triukšmo priklausomybę nuo dažnio. Šiame darbe optimizavimas grindžiamas SNR pripažįstant, kad veiksmingam kompensavimui reikia atsižvelgti į signalo ir triukšmo charakteristikas visame dažnių diapazone.

2. Buvo įrodyta, kad spektrinės kompensacijos bipolinis sužadavimo signalas gali būti pakeistas unipoliniu, o tai gerokai sumažina reikalingos elektronikos sudėtingumą.

3. Buvo pasiūlyta bendroji ultragarso signalų spektrinės optimizacijos sistema, leidžianti kurti įvairias spektrines formas, pritaikytas signalų spektrui arba SNR spektrui.

4. Buvo įrodyta, kad optimalus sužadinimas TRUS reikalauja bent vieno rezonansinio piko ir gretimo spektrinio slėnio aprėpties, kad būtų užtikrintas teisingas atvirkštinis sprendimas.

5. Buvo įrodyta, kad atvirkštinis, filtravimu pagrįstas impulsų suspaudimas, suteikia signalą, lygiavertį delta tipo sužadinimui, leidžiantį suspausti ir filtruoti ilgus signalus. Ši technika gali būti pritaikyta kitoms sujungimo rūšims ir kitoms ultragarsinių matavimų bei vaizdavimo rūšims.

6. Dekompozicijai buvo pasiūlytas liekamasis giluminio mokymosi modelis.

Darbo praktinė vertė

Sukurti metodai leidžia nekontaktiniu būdu matuoti plonų, trapių ar drėgmei jautrių mėginių, tokių kaip biologiniai audiniai, augalų lapai, šilumos izoliacinės medžiagos, popierius, kartonas, filtrai, membranos ar nepolimerizuoti kompozitai, mechaninius parametrus. Derinant oro ultragarsą su TRUS ir naujovišku signalų apdorojimu, įskaitant ilgus programuojamo spektro signalus, impulsų suspaudimą ir spektrinių nuostolių kompensavimą, pagerinama matavimų kokybė naudojant tik vieną keitiklių porą ir įveikiant didelio dažnio slopinimo keliamus sunkumus.

Siūlomo impulsų suspaudimo vertė yra jo rezultatas: gautas atsakas yra lygiavertis didelės energijos plačiajuosčiam impulsui su minimaliais artefaktais, todėl padidinamas SNR ir skiriama geba. Svarbu, kad taikymo sritis apima ne tik TRUS. Tai gali būti naudinga bet kokiems ultragarsiniams matavimams ar vaizdavimo taikymams, kur svarbus didelis SNR ir tikslumas, įskaitant panardinimo ar kontaktinius matavimus.

Svarbiausia, kad pažangios signalo dekompozicijos ir gilaus mokymosi, pagrįsto atspindžių atskyrimo integracija, leidžia efektyviai naudoti ilgus trukmės impulsus atspindinčiose arba sluoksniuotose terpėse, išsaugant spektrinę skiriamąją gebą tiksliam medžiagos savybių nustatymui.

TRUS gali būti išplėstas į 2D arba 3D vaizdų gavimo technologiją, leidžiančią nustatyti medžiagų savybes, pvz., tankį, elastingumo modulį ir storį, kai taikomas erdvinis skenavimas ir fokusuoti spinduliai. Tai atveria galimybes taikyti TRUS ne

tik vieno taško spektroskopijai, bet ir kiekybiniam medžiagų kartografavimui pramonės inspektavime, biomedicininėje diagnostikoje ir žemės ūkio kokybės kontrolei.

Pateikti tyrimų rezultatai buvo pritaikyti šiems projektams:

- Nauji bekontakčiai ultragarso perdavimo metodai plačiajuosčio spektro signalams iš Lietuvos mokslinių tyrimų tarybos, projekto Nr.: S-MIP-23-133 (2023-07-01 - 2025-10-31).
- Ultragarso spektroskopijos jutiklis augalų fiziologijos tyrimams iš Europos regioninės plėtros fondo, projekto Nr.: 01.2.2-LMT-K-718-03-0026 (2020-07-20 – 2023-08-31).

Mokslinė aprobacija

Disertacijos rengimo metu gauti moksliniai rezultatai buvo pateikti 10 mokslinių straipsnių: 3 straipsniai buvo paskelbti tarptautiniuose žurnaluose, įtrauktuose į „Web of Science“ duomenų bazę (Q1), rezultatai buvo pristatyti 8 tarptautinėse konferencijose (3 standiniai ir 5 žodiniai pranešimai) Lisabonoje (Portugalija), Atėnuose (Graikija), Taipėjuje (Taivanas), Kaune (Lietuva), Paderborne (Vokietija) ir Vilniuje (Lietuva). Buvo paskelbti 4 konferencijų pranešimai.

Disertacijos gynimo metu pristatyti rezultatai

1. Buvo sukurta ir eksperimentiškai patvirtinta apibendrinta APWP optimizavimo sistema. Ji leidžia programuoti spektrą pagal išmatuotus SNR profilius, keitiklio atsakus arba tikslinius rezonansus, todėl buvo pasiektas reikšmingas signalo kokybės pagerėjimas: naudojamas –20 dB dažnių diapazonas padidėjo nuo maždaug 480 kHz (įprastiniams impulsiniams ir LFM signalams) iki 670 kHz.

2. Parodyta, kad optimizuotų APWP signalų taikymas oro sujungimo TRUS matavimuose sumažina matavimo paklaidas ir leidžia naudoti tik vieną keitiklių porą plačiame dažnių diapazone, taip pašalinant ankstesnį poreikį oro sujungimo ultragarsiniuose matavimuose naudoti kelis keitiklių rinkinius.

3. Buvo sukurtas ir pasiūlytas sprendimas ilgo išsklaidyto spektro signalų, persidengiančių atspindžiais, problemai spręsti naudojant inversinį filtro suspaudimą, atidarymą ir dekompresiją. Be to, buvo sukurta giliojo mokymosi sistema atspindžiams modeliuoti; pasiektas ~45 dB RE tikslumas ir aukšta koreliacija (~0,9), kai signalo trukmė 50 μ s, o kai signalo trukmė 300 μ s – RE: ~32 dB, koreliacija: ~0,74 tarp prognozuotų ir eksperimentinių signalų.

4. Buvo nustatyta, kad spektrinių nuostolių kompensavimas naudojant ilgus APWP signalus pagerina bendrą TRUS matavimų kokybę stipriai slopinančiomis oro sujungimo sąlygomis, o tikslus rezonansinių parametrų įvertinimas reikalauja, kad sužadavimo spektras apimtų bent vieną pilną rezonansinį piką kartu su gretimu slėniu.

Disertacijos struktūra

Disertacija susideda iš įvado, penkių skyrių, išvadų ir literatūros sąrašo. Disertacijos apimtis – 151 puslapis, 5 lentelės, 53 lygtys, 143 literatūros šaltiniai ir 117 paveikslų.

Pirmoje dalyje aptariama oro jungtimi sujungta RUS, jos taikymas ir svarba NDT matavimams.

Antrajame skyriuje pateikiama SNR pagrįstų APWP signalų plėtra naudojant spektrinį programavimą ir kompensavimo funkcijos plėtrą NC-RUS su veikimo vertinimu.

Trečiajame skyriuje vertinamas signalo spektrinio turinio poveikis storio rezonansinės spektroskopijos matavimų kokybei. Tyrinėjama, kaip keičiasi atvirkštinio sprendimo paklaidos keičiant dažnių diapazoną, kuriame spektriniai nuostoliai kompensuojami atsižvelgiant į rezonansinius maksimumus ir minimumus. Taip pat tiriama, kaip įvairūs siūlomi spektrinio turinio optimizavimo kriterijai veikia atvirkštinio sprendimo kokybę.

Ketvirtajame skyriuje pateikiama atvirkštinio filtro pagrindu sukurta impulsų suspaudimo ir liekamojo giluminio mokymosi metodų plėtra, skirta ilgų išsklaidytų spektro signalų persidengimui spręsti.

Penktajame skyriuje pateikiamas atvirkštinio pulso suspaudimo, atliekamo naudojant filtrą, veikimo vertinimas oro sujungimo ir panardinamuosiuose matavimuose TRUS, naudojant ilgus, išsklaidytą spektrą turinčius signalus.

Problemos analizė

Ultragarinių matavimų tikslumas iš esmės priklauso nuo akustinio sklidimo fizikos, keitiklio charakteristikų ir sužadinimo signalo konstrukcijos sąveikos. Įprasti atvykimo laiko (ToF) metodai užtikrina patikimą atstumo įvertinimą tik tada, kai atskiri atspindžiai yra atskiriami laike. Plonasluoksnėse ar daugiasluoksnėse struktūrose atspindžiai persidengia, todėl tokie metodai tampa neveiksmingi ir reikia taikyti dekonvoliuciją pagrįstas bei rezonansui jautrias technikas, tokias kaip storio rezonansinė ultragarso spektroskopija (TRUS).

Oro sujungimo sistemose papildomą sudėtingumą sukelia stiprus slopinimas, priklausantis nuo dažnio ir atstumo bei aplinkos sąlygų (temperatūros, drėgmės, slėgio) ir turintis įtaką garso greičiui. Kartu su kritinio kampo, artimojo ir tolimojo lauko reiškiniiais šie veiksniai praktiškai riboja matavimus trumpais sklidimo keliais ir žemais dažniais (dažniausiai žemiau 2 MHz). Didėjant atstumui oro slopinimas, spindulio sklaida ir atspindžių nuostoliai mažina jungimo efektyvumą ir iškraipo matuojamą spektrą.

Keitiklio konstrukcija ir sužadinimo strategija lemia pasiekiamą juostos plotį, skiriamąją gebą ir signalo ir triukšmo santykį. Juostos plotį, jautrumą ir perdavimo dažninį atsaką pirmiausia riboja suderinamųjų sluoksnių ir atramos konfigūracija, pjezoelektrinės medžiagos pasirinkimas ir suderinamų medžiagų savybės. Oro sujungimo keitikliai paprastai pasižymi siauresniu juostos plokščiu, nevienodu spektriniu atsaku ir gerokai mažesniu jautrumu nei panardinamieji ar kontaktiniai keitikliai. Vis dėlto nekontaktinė storio rezonansinė ultragarso spektroskopija (NC-RUS) yra patraukli metodika, nes leidžia matuoti plonas, trapias ar drėgmei jautrias

medžiagas jų nepažeidžiant. Analizė rodo, kad daugumos medžiagų pagrindiniai storio rezonansai patenka į maždaug 150 kHz–1,5 MHz dažnių diapazoną, o slopinimas gali siekti nuo –30 dB iki –100 dB, todėl TRUS matavimams būtinas platus juostos plotis ir didelis SNR.

Sužadinant trumpais impulsais pasiekama gera laikinė skiriamoji geba, tačiau prarandama energija ir mažėja signalo ir triukšmo santykis. Išsklaidyto spektro (angl. *spread spectrum*) sužadinimo signalai leidžia atsieti juostos plotį nuo trukmės ir perduoti didesnę energiją nemažinant juostos pločio. Tačiau tam reikia ilgų trukmių signalų, o tai didina pirmojo praeinamo impulso ir dauginių atspindžių persidengimo riziką, ypač tada, kai oro sujungimo keitikliai turi veikti trumpais atstumais. Kai kurios SS signalų rūšys, tokios kaip NLFM ar APWP, sudaro galimybę programuoti spektrinę formą. Ši savybė leidžia kompensuoti transdukcijos spektrinius nuostolius ir efektyviai išplėsti naudingą pralaidumo juostą, todėl galima pagrįstai teigti, kad taikant tokius signalus TRUS matavimų kokybė gali būti iš esmės pagerinta.

APWP signalo kūrimas transdukcijos spektriniams nuostoliams kompensuoti

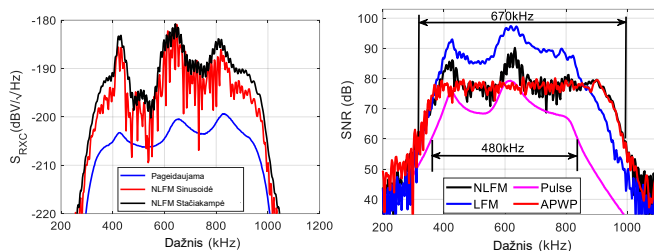
TRUS metodas leidžia vienu metu įvertinti mėginio storį, garso greitį, slopinimą ir tankį, naudojant storio rezonanso modelį [98]. Tačiau oro sujungimo ultragarsas susiduria su esminiais apribojimais – dideliu akustinės varžos nesuderinamumu, stipriu dažnio ir atstumo priklausomumo slopinimu, ribotu juostos pločiu ir mažesniu SNR [99–104]. Nors šiulaikiniai ketvirčio bangos ilgio oro sujungimo keitikliai pasižymi geresniu pralaidumu, praktinį juostos plotį vis dar riboja suderinamųjų sluoksnių medžiagos ir žemas centrinis dažnis, būtinas dėl slopinimo ore.

Tradicinis TRUS dažniausiai taikomas naudojant impulsinius sužadinimo signalus [96, 105, 106]. Jie yra paprasti ir lengvai apdorojami, tačiau dėl siauro impulso sukuriamą mažą energiją ir prastą SNR, o energija koncentruojama žemuose dažniuose, kurie prastai perduodami per oro jungimo keitiklį. SNR galima didinti vidurkinimu, tačiau tai ribota esant kintančioms sąlygoms.

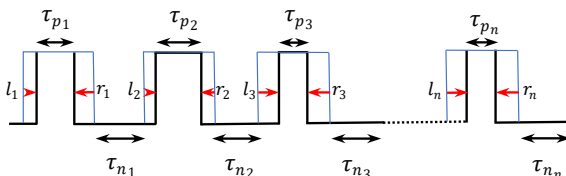
Standartinis sprendimas – kelių keitiklių porų naudojimas siekiant padengti platesnį dažnių diapazoną [96, 105, 106]. Tačiau pakankamą SNR galima pasiekti ir su viena keitiklių pora, jei taikomi išsklaidyto spektro (SS) sužadinimo signalai. Tyrimai rodo, kad LFM ir NLFM signalų spektrinė forma gali būti programuojama nuostoliams kompensuoti [44, 107–108], tačiau bipoliniam sužadinimui reikia sudėtingos elektronikos, o piezokeraminių sistemų triukšmo spektrinė galia nėra tolygi visame dažnių diapazone [109, 110], todėl ankstesniuose metoduose gauti rezultatai nėra optimalūs.

Siekiant šių apribojimų sprendimo, pasiūlytas metodas, pagrįstas unipoliniais APWP sužadinimo signalais, kurie leidžia formuoti spektrinį turinį taip, kad būtų užtikrintas tolygus SNR platesniam juostos pločiui. Skirtingai nuo panardinamųjų ar kontaktinių metodų [51, 53], TRUS taikomas oro sujungimo sąlygomis, o spektrinė optimizacija apima ir perdavimo nuostolius, ir triukšmo spektrą. Pagrindiniai indėliai [36, 50] yra šie: (a) impulsinis sužadinimas pakeistas programuojamos spektro formos SS signalu APWP; (b) žymiai pagerintas SNR be vidurkinimo; (c) spektrinė

optimizacija suteikia tolygų SNR pasirinktame diapazone (1 pav.); (d) unipoliniai APWP signalai supaprastina elektroniką ir išlaiko valdomą spektrą (2 pav.).



118 pav. Priimto signalo amplitudės spektro (kairėje) ir SNR (dešinėje) palyginimas, kai oro jungtimi sujungtas keitiklis sužadinamas LFM (raudona) ir SNR nuostolius kompensuojančiu sinusoidiniu NLFM (juoda), stačiakampiu NLFM (raudona) ir APWP (juoda) signalu; pageidaujamas atsakas – mėlyna kreivė



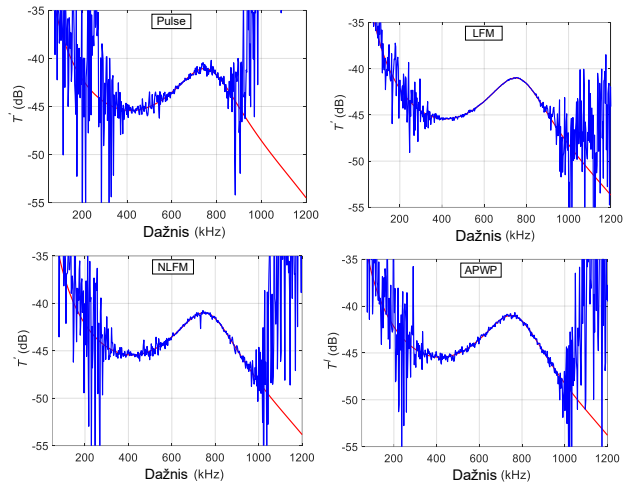
119 pav. Vienpolinis sužadinimo signalas APWP

Du 650 kHz centrinio dažnio keitikliai (sukurti ir pagaminti CSIC, Ispanija) buvo pastatyti 30 mm atstumu.

Plokščio SNR veikimo vertinimas TRUS matavimuose

Siekiant įvertinti siūlomo metodo veikimą, buvo atlikta skaitmeninė imitacija ir eksperimentinis tyrimas. Imitacijai pasirinkti *Vitis vinifera* lapai dėl jų kintamų mechaninių savybių ir sudėtingos struktūros, todėl realių matavimų stabiliai atlikti neįmanoma. Eksperimentinis įvertinimas atliktas naudojant polikarbonato plokštę, kurios storis 2,045 mm, o išilginis garso greitis apie 2240 m/s. Abiem atvejais matavimų tikslumas įvertintas taikant storio rezonanso modelį. Siekiant atskleisti realias sąlygas prie imituotų signalų pridėtas Gauso triukšmas.

Palyginti keturi sužadinimo signalai: impulsinis, LFM, NLFM ir APWP. Impulsinis sužadinimas pasižymi siauru juostos pločiu ir mažu energijos kiekiu, todėl SNR išlieka mažiausias. LFM suteikia didžiausią bendrą energiją, tačiau SNR pasiskirstymas yra nevienodas, o didžiausia energija koncentruojama į žemus dažnius. NLFM užtikrina platesnį efektyvų juostos plotį, o APWP suteikia plačiausią ir tolygiausią spektrinę aprėptį, nes leidžia programuoti spektrinę formą ir kompensuoti keitiklio transdukcijos nuostolius (120 pav.).



120 pav. Perdavimo dydis *Vitis Vinifera* lapui esant skirtingiems sužadavimo signalams: impulsui (viršuje kairėje), LFM (viršuje dešinėje), NLFM (apačioje kairėje) ir APWP (apačioje dešinėje)

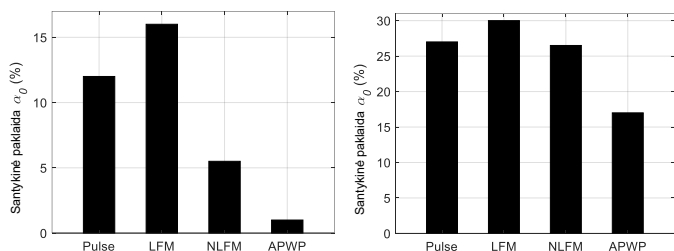
Eksperimentiniuose polikarbonato matavimuose visi keturi sužadavimo tipai generavo rezonansinius atsakus, tačiau jų spektrinė aprėptis skyrėsi. Impulsinis ir LFM signalai parodė ribotą spektrinį sutapimą su mėginio rezonanso juosta, todėl parametrai (storis, greitis, slopinimas, tankis) įvertinami su didesnėmis klaidomis. NLFM parodė platesnę aprėptį, tačiau geriausią spektro atitikimą užtikrina APWP signalas, nes jis efektyviausiai kompensuoja spektrinius nuostolius ir suderina energijos paskirstymą visame dominančiame dažnių ruože.

Galutiniai rezultatai rodo, kad APWP sužadimas užtikrina didžiausią TRUS matavimų tikslumą ir imituojant, ir eksperimente. Toks signalas leidžia pasiekti platesnį pralaidumo ruožą, didesnę SNR ir tikslesnę parametru įvertinimą nereikalaujant vidurkinimo ar kelių keitiklių porų. Kiti signalai, ypač impulsinis ir LFM, dėl ribotos spektrinės aprėpties yra mažiau tinkami tiksliai analizuoti storio rezonansą.

Taigi, APWP signalo naudojimas yra veiksmingiausia priemonė spektriniams nuostoliams kompensuoti, rezonanso pikų aprėpčiai ir parametru įvertinimo klaidoms mažinti TRUS matavimuose.

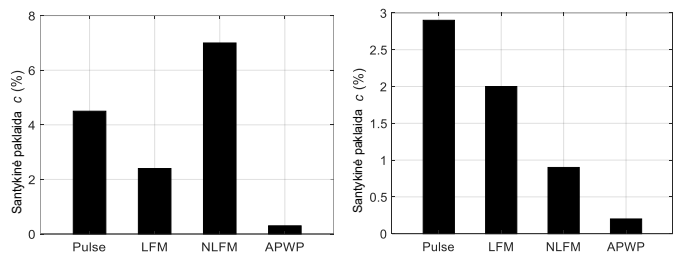
Mėginio parametru įvertinimo našumas

Slopinimo a įvertinimo paklaidos pateiktos 121–124 paveikslė. Kairėje pusėje pateikti *Vitis vinifera*, o dešinėje – polikarbonato mėginiai.



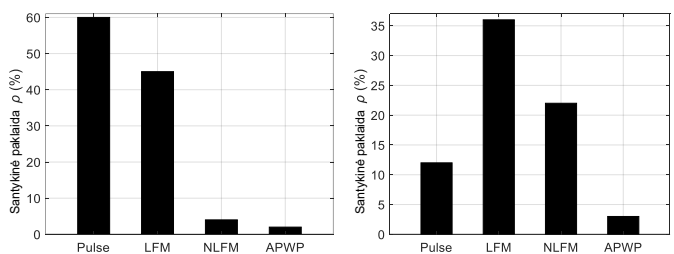
121 pav. Santykinė paklaida a : *Vitis vinifera* (kairėje) ir polikarbonato (dešinėje)

Galima matyti, kad didžiausios paklaidos yra LFM sužadavimo atveju (16 %), po to eina impulsas (12 %), NLFM (5,5 %), o mažiausios paklaidos yra naudojant SNR optimizuotą APWP (1 %) *Vitis vinifera* atveju. Reikia paminėti, kad eksperimentiniai rezultatai naudojant polikarbonatą taip pat yra geresni APWP atveju (17 %).

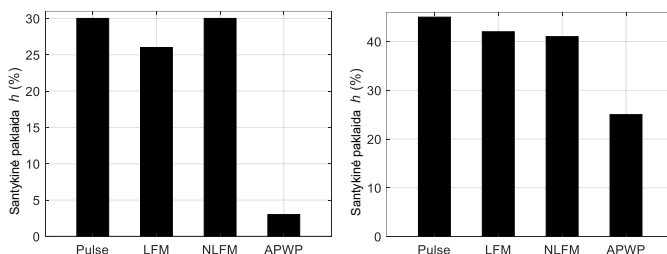


122 pav. Santykinė paklaida c : *Vitis vinifera* (kairėje) ir polikarbonato (dešinėje)

Santykinės c įvertinimo nuokrypio paklaidos yra didžiausios NLFM atveju (7 %), po to eina impulsas (4,5 %) ir LFM (2,4 %), o APWP rezultatai yra mažiausi – tik 0,3 %. Nuokrypio paklaidos polikarbonato atveju yra didžiausios impulsui (2,9 %) ir mažiausios APWP (0,2 %).



123 pav. Santykinė paklaida r *Vitis vinifera* (kairėje) ir polikarbonato (dešinėje) atveju



124 pav. Santykinė paklaida h *Vitis vinifera* (kairėje) ir polikarbonato (dešinėje) atveju

Abiejų atvejų r ir h paklaidos taip pat mažiausios APWP atveju. Todėl eksperimentiškai vertinant galima daryti išvadą, kad spektrinė nuostolių kompensacija su SNR išlyginimu sumažina TRUS matavimų paklaidas.

Išvados

Metodika buvo išbandyta tiek oro sujungimo, tiek panardinamajame ultragarse. Siekiant užtikrinti aukštą SNR pralaidumą juostoje, buvo pasiūlytas SNR optimizuotas APWP signalas su programuojama spektrine forma. Parodyta, kad išsklaidytos spektro (angl. *spread spectrum*) sekos leidžia pasiekti tinkamą SNR platesniame negu keitiklio pralaidumo ruože, kas neįmanoma naudojant impulsus. APWP signalas ties -20 dB užtikrina 670 kHz pralaidumo juostą, o impulsų -20 dB juosta siekia tik 480 kHz.

Nustatyta, kad vienpoliai sužadinimo signalai suteikia pakankamas spektro formavimo galimybes, o jų naudojimas leidžia supaprastinti sužadinimo elektroniką.

Siūlomas metodas įvertintas taikant TRUS modeliavimą (*Vitis vinifera*) ir eksperimentą (polikarbonato plokštė). Modeliavimo ir eksperimentiniai rezultatai parodė, kad spektrinės kompensacijos signalai (APWP) užtikrina mažiausias visų parametų įvertinimo paklaidas.

Santykinės tankio įvertinimo paklaidos buvo mažiausios naudojant SNR optimizuotą APWP (1 %) *Vitis vinifera* imitacijos atveju. Didžiausios paklaidos gautos taikant LFM (16 %), impulsą (12 %) ir NLFM (5,5 %). Eksperimentiniu būdu nustatyta, kad polikarbonato plokštės matavimuose mažiausios tankio paklaidos taip pat gautos APWP (17 %), o didžiausios – LFM (30 %), impulsų (27 %) ir NLFM (27 %).

Santykinės garso greičio įvertinimo paklaidos buvo mažiausios naudojant SNR optimizuotą APWP (0,3 %) *Vitis vinifera* imitacijos atveju. Didžiausios paklaidos buvo gautos su NLFM (7 %), po to su impulsu (4,5 %) ir LFM (2,4 %). Eksperimentiniuose matavimuose mažiausia greičio paklaida taip pat gauta naudojant APWP (0,2 %), o didžiausios – impulsui (2,9 %), LFM (2 %) ir NLFM (1,9 %).

Tankio (ρ) ir storio (h) įvertinimo paklaidos abiem atvejais taip pat buvo mažiausios naudojant APWP signalą.

Spektrinio turinio poveikio tyrimas bekontaktėje rezonansinėje ultragarso spektroskopijoje

APWP signalai buvo pasiūlyti transdukcijos nuostoliams kompensuoti, o ankstesniame skyriuje parodyta, kad SNR išlyginimas leidžia išplėsti efektyvų pralaidumo ruožą. Vis dėlto TRUS optimizavimo kriterijai išliko neaiškūs. Šiame skyriuje nagrinėjami skirtingi spektrinės formos programavimo kriterijai ir analizuojama, kokia spektrinė aprėptis yra reikalinga tam, kad būtų sumažintos matavimų paklaidos atsižvelgiant į mėginio rezonanso viršūnes ir slėnius.

Buvo išbandyti šeši APWP optimizavimo kriterijai: (a) SNR lygumo užtikrinimas; (b) LFM SNR gerinimas linijine skale; (c) LFM SNR gerinimas su papildomu SNR svoriu; (d) LFM energijos didinimas dB skale; (e) NLFM SNR gerinimas linijine skale; (f) NLFM $1/\text{SNR}$ gerinimas dB skale. Visi kriterijai optimizuoti tik pasirinktoje dažnių srityje $f_{\min}-f_{\max}$. Kompensavimo funkcija paremta dabartinės APWP signalo versijos AC SNR atsaku.

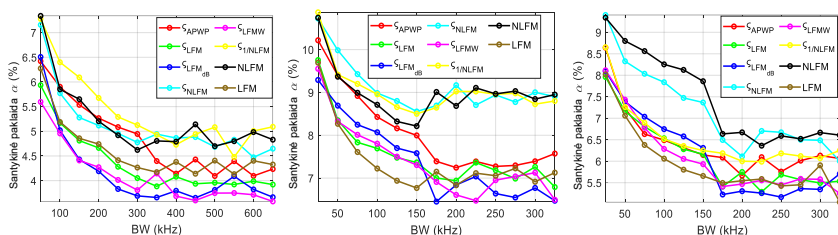
Kriterijai nuo (a) iki (d) leidžia gerinti spektrą lyginant su LFM, o (e) ir (f) – lyginant su NLFM, priklausomai nuo to ar energija turi būti didinama, ar mažinama tam tikruose dažniuose. Pagrindinis tikslas – perkelti energiją į tas pralaidumo juostos sritis, kuriose transdukcijos nuostoliai yra didžiausi, taip užtikrinant vienodesnį SNR ir didesnę efektyvų juostos plotį.

Eksperimentiškai metodas įvertintas naudojant ploną 2,05 mm polikarbonato mėginį ir du 650 kHz oro sujungimo keitiklius, išdėstytus 35 mm atstumu perdavimo režime. Sužadinimui naudota 35 V unipolinė APWP seka (100 μ s trukmės), o duomenų rinkimas ir visi skaičiavimai (atvirkštinis sprendimas, optimizavimas, vertinimas) atlikti MATLAB aplinkoje. Aplinkos poveikiui sumažinti taikyta oro parametrų kompensacija [43], o mėginio ir kalibravimo matavimai automatiškai atlikti po 100 kartų naudojant servovariklio padavimo mechanizmą.

Sudaryti trys eksperimentų rinkiniai: (a) siaura juosta aplink pagrindinį rezonanso piką (~620 kHz); (b) juostos plėtimas aukštyrų nuo rezonanso; (c) juostos plėtimas žemyn nuo rezonanso. Juostos plotis didintas 25 kHz žingsniu iki stabdymo juostos, o kiekviena spektrinė kombinacija buvo panaudota TRUS parametrų įvertinimui.

Rezultatai

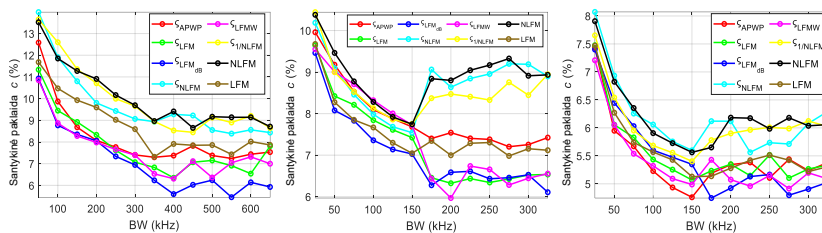
Kiekvieno eksperimento paklaidos buvo sujungtos ir įvertintos keturiems parametrams: silpnumui α_0 , greičiui c , tankiui ρ ir storiui h . Nuokrypio paklaidos apskaičiuotos palyginant matavimų vidurkį su žinomomis polikarbonato reikšmėmis ($\alpha_0 = 43,5$ Np/m, $c = 2240$ m/s, $h = 2,045$ mm, $\rho = 1193,6$ kg/m³). Rezultatai rodo, kad visais trimis spektrinės aprėpties scenarijais – kai spektras koncentruojamas ties rezonanso piku, plečiamas aukštesnių arba žemesnių dažnių link – siaura pralaidumo juosta lemia dideles paklaidas (125–127 pav.). Kai spektrinis turinys apima pagrindinį rezonanso piką ir išsiplečia pusiaukelėje iki artimiausio slėnio, paklaidos reikšmingai sumažėja, o tolesnis juostos plėtimas jau neduoda esminio pagerėjimo.



125 pav. Polikarbonato mėginio silpnėjimo nuokrypio klaidos, kai sužadinimo aprėptis yra sutelkta aplink rezonanso piką (kairėje), plečiasi nuo rezonanso piko link aukštesnių dažnių (viduryje) ir plečiasi nuo rezonanso piko link žemesnių dažnių (dešinėje)

Geriausi rezultatai pasiekti naudojant LFM ir LFM pagrįstą APWP, taikant kriterijus ζ_{LFMdB} (kur pirmenybė teikiama bet kokiam energijos pagerinimui, palyginti su LFM). Visi NLFM pagrįsti signalai pasirodė blogiausiai. Tokių rezultatų buvo galima tikėtis, nes silpnėjimo įtaka yra tik apie piką. SNR pikas yra maždaug toje pačioje vietoje kaip ir bandinio rezonanso pikas. Todėl pakanka naudoti LFM signalą su pakankama spektrine aprėptimi (apimantį piką ir pusę kelio iki slėnio). LFM pagrįstas APWP veikia geriau, bet skirtumas yra nežymus.

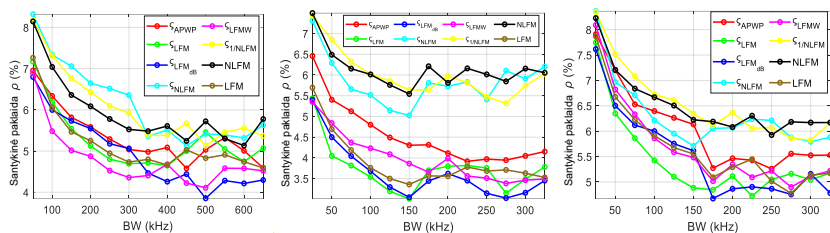
Eksperimentiniai NC-RUS naudojant ultragarso greičio įvertinimo paklaidos rezultatai pateikti 126 pav.



126 pav. Ultragarso greičio nuokrypio paklaidos polikarbonato mėginyje, kai sužadinimo aprėptis yra (kairėje) sutelkta aplink rezonanso piką, (viduryje) plečiasi nuo rezonanso piko link aukštesnių dažnių ir (dešinėje) plečiasi nuo rezonanso piko link žemesnių dažnių

Tendencija yra tokia pati kaip ir ultragarso greičio atveju: kai rezonanso pikas ir pusiaukelė iki slėnio (pusiaukelė yra 300 kHz kairėje ir 150 kHz centre bei dešinėje) yra padengti, paklaidos sumažėjimas nėra toks stiprus. Geriausi rezultatai pasiekti naudojant LFM pagrįstą APWP, taikant kriterijus ζ_{LFMdB} , ζ_{LFM} (kur teikiama pirmenybė bet kokiam energijos pagerinimui, palyginti su LFM) ir ζ_{LFMW} (kur energija bandoma pagerinti, palyginti su LFM tik tų dažnių atveju, kur spektriniai nuostoliai yra dideli). Visi NLFM pagrįsti signalai pasirodė blogiausiai. Rezultatas yra tikėtinas, nes greičio įtaka yra daugiausia apie piką. Tačiau slėnis taip pat turi tam tikrą reikšmę, todėl LFM pasiekė blogesnius rezultatus (nepakankamas SNR slėnyje). Šiek tiek geresni rezultatai pasiekiami, kai padengiami žemesni dažniai slėnyje (10 pav. dešinėje).

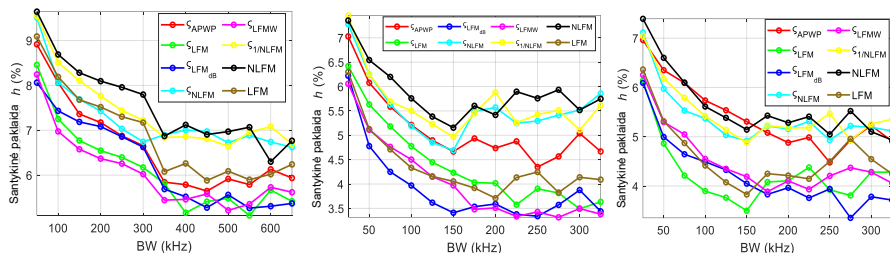
Polikarbonato mėginio tankio paklaidos parodytos 127 pav.



127 pav. Polikarbonato mėginio tankio nuokrypio paklaidos, kai sužadavimo aprėptis yra (kairėje) sutelkta aplink rezonanso piką, (viduryje) plečiasi nuo rezonanso piko link aukštesnių dažnių ir (dešinėje) plečiasi nuo rezonanso piko link žemesnių dažnių

Bendroji tendencija yra ta pati, tik paklaidos mažėja iki slėnio (600 kHz kairėje ir 300 kHz viduryje ir dešinėje), o už šio taško paklaidos mažėja nežymiai. Galima teigti, kad reikšmingiausias yra LFM pagrįstas APWP, naudojantis kriterijus ζ_{LFMdB} (bet koks energijos pagerėjimas, palyginti su LFM, yra palankus). Visi NLFM pagrįsti signalai pasirodė blogiausiai. Rezultatą galima paaiškinti: nors viršūnė ir slėnis yra veikiami tankio, tankio įvertinimas yra pagrįstas impedansu, todėl jis yra susijęs su greičiu (poveikis aplink viršūnę). Tačiau slėnis taip pat turi tam tikrą reikšmę, todėl LFM pasirodė blogiau.

Nuokrypio klaidos matuojant storį pateiktos 128 paveiksle.



128 pav. Polikarbonato mėginio storio nuokrypio paklaidos, kai sužadavimo aprėptis yra (kairėje) sutelkta aplink rezonanso piką, (viduryje) plečiasi nuo rezonanso piko link aukštesnių dažnių ir (dešinėje) plečiasi nuo rezonanso piko link žemesnių dažnių

Paklaidų analizė parodė aiškią tendenciją: jos reikšmingai sumažėja iki rezonanso slėnio pasiekimo, o vieno slėnio aprėptis netgi duoda geresnių rezultatų (apie 3,5 %) nei abiejų slėnių aprėptis (apie 5,5 %). Geriausia tikslumo zona pasiekama padengus aukštesnių dažnių slėnį, esantį iškart už rezonanso viršūnės. Optimaliausi rezultatai gauti naudojant LFM pagrįstą APWP su kriterijais ζ_{LFMdB} ir ζ_{LFMW} , kurie dB skalėje pirmenybę teikia energijos padidėjimui (bendram arba tik ten, kur nuostoliai dideli). Nors ζ_{LFMW} generuoja lygiausią spektrą, jo veikimas prastesnis, nes optimizavimas (baudžia) net ir už SNR pagerėjimą. Bendra išvada tokia: tikslumui svarbiausia, kad pralaidumo juosta apimtų rezonanso piką ir bent vieną slėnį; tolesnis juostos plėtimas turi nedidelį poveikį, tačiau jokių būdu nepablogina rezultatų. Todėl platesnės juostos signalai, kurie pagerina SNR pasirinktoje srityje, visada mažina

paklaidas, o geriausią veikimą užtikrina LFM pagrįstas APWP, optimizuotas pagal ζ_{LFMdB} kriterijų.

Išvados

Spektrinės aprėpties klaidų poveikio matavimo vertinimas parodė, kad pakanka, jei spektrinis turinys apima mėginio rezonanso piką ir vieną slėnį. Mažiausios tankio (3 %) ir storio (3,5 %) klaidos gaunamos tuo atveju, kai spektrinis turinys apima rezonanso slėnį, esantį aukštų dažnių dalyje. Kai spektrinis turinys apima žemesnių dažnių slėnį, geriausias rezultatas gaunamas greičio (5 %). Įtraukus abu slėnius, gaunama mažiausia silpnėjimo paklaida (3,8 %). Kompromisas, tenkinantis visus parametrus yra tada, kai apimama visa -20 dB pralaidumo juosta – pastebimas našumo sumažėjimas.

Buvo įrodyta, kad APWP signalai yra geresni už signalus, iš kurių jie yra gaunami. LFM pagrįsti APWP, naudojant ζ_{LFMdB} kriterijų, duoda geriausius rezultatus. LFM signalai su plačia spektrine aprėptimi taip pat gali būti naudojami esant kelių procentų paklaidai.

Sprendimas ilgų signalų persidengimo atveju

Naudojant suderintą filtrą SS signalams suspausti atsiranda koreliacijos šoniniai sparnai, kurie persidengia su mėginio aidu ir pablogina silpnėjimo bei tankio įvertinimą. Kuo ilgesnis SS signalas, tuo labiau jo šoniniai žiedai iškreipia TRUS atsaką.

Šiai problemai spręsti buvo pasiūlyta naudoti impulsų suspaudimą atvirkštinio filtru. Šis metodas leidžia gauti sužadavimo signalą be šoninių žiedų, išlaikant geresnį dažninį atsaką ypač aukštesnių dažnių srityje. Gauti suspausti signalai yra artimi skleisto juostos pločio *Dirac* impulsui, todėl tinka tiksliai TRUS analizei. Tačiau trūkumas tas, kad nebėra tiesioginės galimybės programuoti sužadavimo spektro nuostoliams kompensuoti – impulsų suspaudimas iškreipia plokščią spektrinę formą.

Kad šis metodas būtų taikomas TRUS, buvo pasiūlytas atidarymo ir dekompresijos derinys. Pirmiausia taikomas atidarymas – jis leidžia atskirti kalibravimo ir mėginio signalus, nes šoniniai žiedai nebesusidaro. Po vartų signalas atkuriamas dekompresijos metodu, grąžinant jį atgal į laiko sritį, bet jau be persidengimų. Taip gaunamas signalas, tinkamas TRUS parametrų įvertinti išlaikant aukštą SNR ir tikslų perdavimo piko dažnių atkūrimą.

Našumo vertinimas

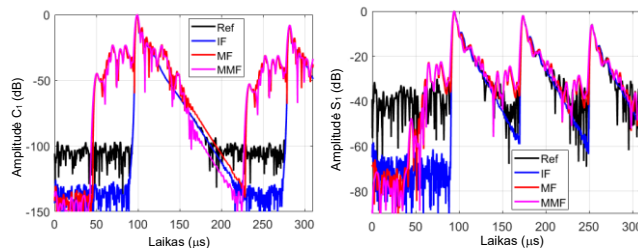
Našumui vertinti buvo naudojami trys skirtingi signalų rinkiniai: a) LFM (250 kHz–1,1 MHz), b) APWP (380 kHz–900 kHz), optimizuotas plokščiam spektrui esant -20 dB, c) APWP (380 kHz–900 kHz), suderintas su keitiklio AC atsakymu esant -20 dB. Buvo tirta dešimt skirtingų impulsų suspaudimo tipų: a) neapdorotas signalas (pažymėtas: *Raw*); b) impulsų suspaudimas naudojant atvirkštinį filtrą (pažymėtas: IF); c) impulsų suspaudimas naudojant suderintą filtrą (pažymėtas: MF); d) impulsų

suspaudimas naudojant nesuderintą filtrą (pažymėtas: MMF); e) impulsų suspaudimas naudojant *Wiener* filtrą (pažymėtas: WF); f) neapdorotas signalas su sumažintais persidengiančiais atspindžiais (be vartų); g) impulsų suspaudimas naudojant atvirkštinį filtrą ir dekompresiją (pažymėtas: IFd); h) impulsų suspaudimas naudojant suderintą filtrą ir dekompresiją (pažymėtas: MFd); i) impulsų suspaudimas naudojant nesuderintą filtrą ir dekompresiją (pažymėtas: MMFd); m) impulsų suspaudimas naudojant *Wiener* filtrą ir dekompresiją (pažymėtas: WFd). Atveju (n) suspausti atvirkštinio filtro signalai pirmiausia buvo filtruojami, o tada dekompresuojami, taip atimant žaliąjį signalą.

Imitacijos rezultatai

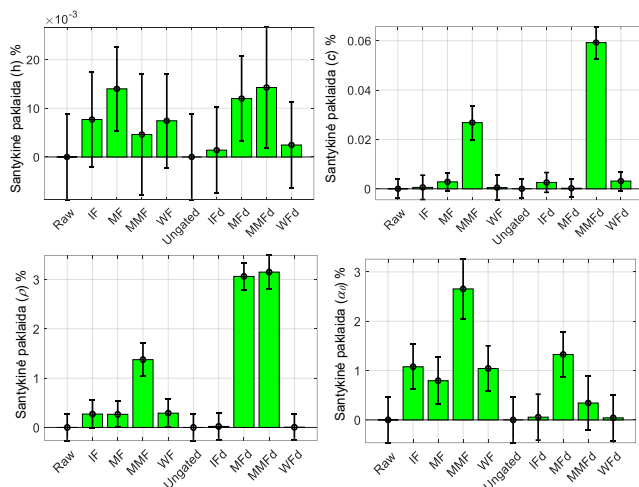
Impulso suspaudimo buvo analizuoti trys sužadinimo signalų rinkiniai – LFM, APWP su plokščiu -20 dB spektru ir APWP su -20 dB suderinimu su keitiklio AC atsaku. Iš visų išbandytų impulsų suspaudimo metodų paaiškėjo, kad tik atvirkštinis filtras (IF) ir *Wiener* filtras (WF) yra tinkami ilgos trukmės išsklaidytų spektro signalams suspausti, nes jie neformuoja šoninių žiedų, kurie persidengtų su mėginio signalo pabaiga. Likę metodai (MF, MMF ir jų dekompresijos variantai), ypač ilgų signalų atveju, sukuria ryškius koreliacijos artefaktus, gerokai padidinančius parametrų paklaidas.

Ilgėjant signalo trukmei, IF ir WF leidžia išlaikyti švarų laiko srityje atsaką, o tai pavaizduota 129 pav., kur pateikiamos Hilberto transformacijos $50\ \mu\text{s}$, $100\ \mu\text{s}$ ir $350\ \mu\text{s}$ LFM atvejams. $50\ \mu\text{s}$ signalo atveju šoniniai sparneliai dar nepersidengia, tačiau $100\ \mu\text{s}$ ir $350\ \mu\text{s}$ atvejais MF ir MMF metodais jie visiškai uždengia mėginio pabaigą, todėl NC-RUS tampa netinkamas be papildomo filtravimo.



129 pav. Kalibravimo (kairėje) ir pavyzdžio (dešinėje) signalų Hilberto apgaubos, kai LFM sužadinimo trukmė yra $50\ \mu\text{s}$

SNR skirtumas tarp trumpų ($10\ \mu\text{s}$) ir ilgų ($350\ \mu\text{s}$) signalų yra didelis – trumpų signalų tikslumą riboja triukšmas, o ilgų – šoniniai žiedai. Tai patvirtina 130 pav., kur pateiktos $10\ \mu\text{s}$ $20\ \text{V}$ LFM nuokrypio paklaidos: visų metodų rezultatai panašūs, nes SNR yra žemas. O ilgų signalų ($350\ \mu\text{s}$) atveju MF, MMF ir jų dekompresijos metodai duoda kelis kartus didesnių paklaidų dėl persidengiančių atgarsio komponentų.



130 pav. Nuokrypio paklaidos 10 ms 20 V LFM

Siekiant sumažinti šiuos apribojimus buvo analizuoti spektriškai suformuoti APWP signalai, kurie leido padidinti energiją dažniuose, kur keitiklio jautrumas yra mažiausias.

Apibendrinant, naudoti galima tik tuos metodus, kurie nesukuria šoninių sparnelių ir išlaiko aiškią signalo pabaigą. IF ir WF yra vieninteliai suspaudimo metodai, tinkami NC-RUS, o spektriškai kompensuotas APWP suteikia mažiausias sisteminės ir atsitiktines paklaidas ir imituojant, ir eksperimente.

Giliųjų neuroninių tinklų metodas sprendimams, kurie sutampa

Oro jungiamojo ultragarso matavimai per perdavimą reikalauja kalibravimo ir mėginių atsako įvertinimo, kuriame svarbus dažninės srities perdavimo santykis. Tradiciškai naudojami trumpi impulsiniai signalai suteikia gerą laikinę skiriamąją gebą, tačiau jų juostos plotis ir SNR yra riboti. Dėl to vis dažniau naudojami platūs spektro signalai, tokie kaip LFM, kurie paskirsto energiją platesniame dažnių diapazone ir leidžia pasiekti geresnę spektrinę aprėptį.

Tačiau oro jungtyje keitiklių atstumai yra riboti, todėl ilgesnės trukmės signalai dažnai sukelia sudėtingus persidengiančius atspindžius, kurių neįmanoma atskirti nei signalo langavimu, nei suderintu, nei atvirkštiniu filtravimu. Šie tradiciniai metodai reikalauja žinomo sistemos impulsinio atsako, kuris oro jungiamose sistemose paprastai nėra tiksliai žinomas, todėl jų taikymas tampa nepatikimas.

Dėl šių priežasčių pradėti taikyti giliaji neuroniniai tinklai, kurie geba mokytis iš eksperimentinių duomenų ir rekonstruoti signalus, net jei atspindžiai stipriai persidengia. Siūlomas metodas remiasi 1D konvoliuciniu liekamuoju UNet (Res-UNet1D) tinklu, sukurtu specialiai oro jungiamų ultragarsinių bangų formų rekonstrukcijai. Modelis buvo mokomas trumpų LFM signalų diapazone (10–50 μ s), kuriame atspindžiai natūraliai atskirti ir galima tiksliai išskirti pirmąjį atspindį. Šis

atspindys laikomas tiksliniu mokymo rezultatu, o įvestis – eksperimentiniu būdu išmatuota sužadavimo įtampa.

Prieš mokymą visi signalai buvo normalizuojami ir filtruojami, kad būtų pašalintas nepageidaujamas triukšmas. Res-UNet1D architektūra su dilatacinėmis konvoliucijomis leidžia efektyviai užfiksuoti ilgalaikius laiko priklausomumus ir atkurti aiškų pirmąjį atspindį net tada, kai ilgų trukmių signalai turi stiprų persidengimą. Modelis buvo testuotas 100–350 μ s trukmės LFM signalais, kurie nebuvo mokymo dalis ir parodė gebėjimą generuoti švarius, neperdengtus pirmuosius atspindžius išlaikydamas tiek laiko, tiek dažnio charakteristikų nuoseklumą. 50 μ s trukmės rezultatai parodyti 131 pav.



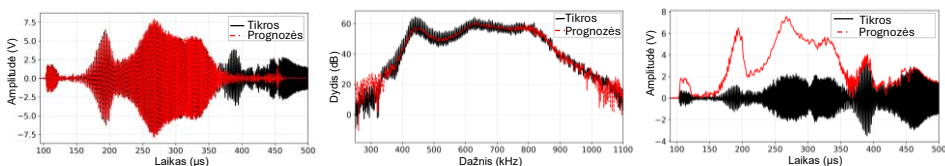
131 pav. Laikinis (kairėje), spektrinis (viduryje) ir liekamasis (dešinėje) atsakas 50 μ s, kuris buvo mokymo dalis

Šis metodas suteikia naują būdą spręsti persidengiančių signalų problemą oro jungiamame ultragarse, kur tradiciniai filtravimo metodai yra riboti. Res-UNet1D leidžia rekonstruoti kokybiškus bangų signalus iš sudėtingų eksperimentinių duomenų ir turi potencialo būti taikomas tolesniuose TRUS parametrų įvertinimo etapuose.

Res-UNet1D modelio rezultatai ilgos trukmės signalams

Res-UNet1D modelis buvo sukurtas siekiant prognozuoti ultragarsinius signalus su stipriu atspindžių persidengimu, o jo gebėjimas apibendrinti buvo vertinamas naudojant ilgesnės trukmės LFM signalus, kurie nebuvo mokymo dalis. Modelio taikymas 250 μ s signalui parodė, kad, nors liekamoji paklaida yra didesnė nei trumpesnių trukmių atveju, prognozės išlieka priimtinos atsižvelgiant į triukšmo lygį ir tai, jog modelis buvo mokytas tik su 10–50 μ s trukmės signalais.

Ilgiausio, 350 μ s signalo rekonstrukcija taip pat patvirtino metodo stabilumą – liekamieji iškraipymai išlieka maži pirmojo atspindžio regione, o laiko ir dažnio sričių palyginimas yra pakankamai tikslus. Tai iliustruojama 132 pav.



132 pav. Laikinis (viršuje), spektrinis (centre) ir likutinis (dešinėje) 350 μ s atsakas, kuris buvo mokymo dalis

Šie rezultatai rodo giluminio mokymosi metodo efektyvumą modeliuojant persidengiančius atspindžius oro jungiamojo ultragarso sistemoje ir patvirtina, kad Res-UNet1D gali išmokti rekonstruoti pirmąjį atspindį net iš labai ilgos trukmės sudėtingų signalų. Šiame tyrime metodas buvo taikytas tik kalibravimo signalams, kurie yra paprastesni nei mėginio perdavimo atsakai, todėl tolesni darbai turėtų apimti metodo išplėtimą mėginių matavimams bei skirtingų signalų tipų (impulsinių, NLFM) panaudojimą. Nepaisant šių ribojimų, gauti rezultatai rodo, kad siūlomas požiūris yra perspektyvus ir gali būti toliau tobulinamas siekiant platesnio taikymo ultragarso signalų rekonstrukcijoje.

Išvados

Dėl netiesiškumų ore ir impedanso nesuderinamumo NC-RUS vis labiau didėja ilgos trukmės plitimo spektro signalų paklausa, SS signalai išplečia analizės metu prieinamą dažnių juostą ir pasiekia aukštą SNR. Dėl riboto atstumo tarp keitiklių ir ilgos SS signalų trukmės, ultragarso matavimų atspindžiai sutampa, todėl sunku naudoti signalus apdorojimui suspaudimą atvirkštiniu filtru.

Sutampančių signalų atspindžių sprendimui buvo naudojami du skirtingi metodai: giluminiai neuroniniai tinklai, pagrįsti sutampančiais atspindžiais, ir atvirkštinis filtras, pagrįstas impulsų suspaudimu sutampantiems atspindžiams atkurti.

Res-UNet1D pagrįstas gilusis mokymasis buvo pasiūlytas ilgalaikiams atspindžių persidengimams *chirp* signaluose, naudojamuose oro jungiamam ultragarsui. Mokymo procesas buvo vykdomas naudojant trumpalaikius signalus, o testavimas buvo atliekamas tiek trumpalaikiams, tiek ilgalaikiams signalams. Gilusis mokymasis buvo išbandytas tik kalibravimo signalo modeliavimui; vis dėlto literatūroje analogiškas metodas iki šiol nebuvo taikytas, todėl ateities tyrimai bus sutelkti į perdavimo atsaką arba NC-RUS matavimų pavyzdžius.

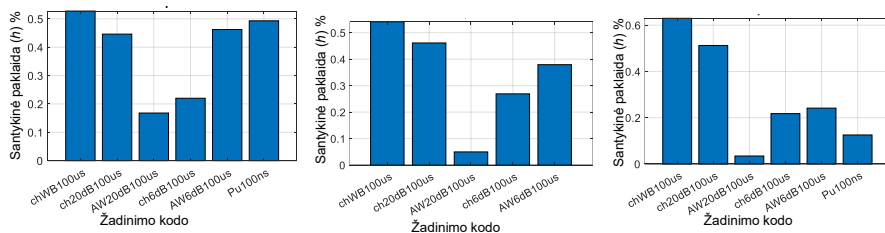
NC-RUS buvo pasiūlytas ir išbandytas atvirkštinio filtro pagrįstas impulsų suspaudimo metodas, naudojant modeliavimą ir eksperimentinę analizę. Naudojant atvirkštinį filtrą su suspausto signalo vartų valdymu ir dekompresija, galima reikšmingai sumažinti įvertinimo paklaidas.

Našumo vertinimas storio rezonansinės ultragarso spektroskopijos atveju

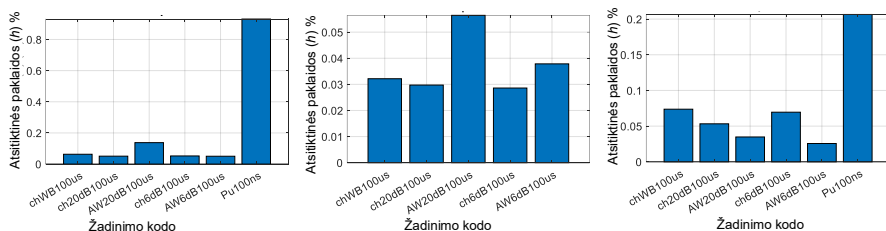
Turint visą siūlomą signalų optimizavimo ir apdorojimo metodiką, buvo įvertintas TRUS veikimas naudojant ilgus, persidengiančius plitimo spektro signalus. Signalai buvo suspaudžiami atvirkštiniu filtru, vartomi, dekompresuojami ir tik tada pateikiami TRUS apdorojimui, kaip aprašyta V skyriuje. Tyrimui naudotas 2 mm polikarbonato mėginys, kurio pagrindinis storio rezonansas yra ties 560 kHz, o etaloniniai parametrai (storis, greitis, tankis ir slopinimas) nustatyti iš kalibravimo matavimų. Buvo atlikti du eksperimentiniai rinkiniai: oro jungties NC-RUS ir panardinimo TRUS, siekiant palyginti veikimą skirtingose sujungimo terpėse. Oro jungties konfigūracijoje naudoti plačiajuosčiai 650 kHz keitikliai, o panardinimo atveju – kompoziciniai 2 MHz keitikliai.

Oro jungiamų NC-RUS matavimų rezultatai

Oro jungiamų NC-RUS matavimų analizė parodė, kad APWP signalai su spektrinių nuostolių kompensavimu užtikrina geriausią TRUS veikimą. Plačiajuostis APWP, optimizuotas -20 dB pralaidumo juostoje, suteikė mažiausias sisteminės paklaidas didelio SNR sąlygomis, o siaurajuostis APWP su -6 dB kompensavimu pasižymėjo mažiausia atsitiktine paklaida. Visuose matavimuose APWP SNR buvo pagerintas pereinamuosiuose dažnių regionuose, o tai leido išplėsti efektyvų pralaidumo diapazoną, viršijantį LFM ir impulsinių signalų galimybes. Esant mažesniai sužadino lygiui (20 V), laiko pagrindu apskaičiuojami parametrai (storis ir greitis) išliko tikslūs, o amplitudės priklausomi parametrai (tankis ir slopinimas) jautriau reagavo į SNR sumažėjimą. Nuokrypio ir atsitiktinių paklaidų įvertinimai (133 ir 134 pav.) rodo, kad pagrindinis matavimo neapibrėžtumo šaltinis yra sisteminės paklaidos, o atsitiktinės paklaidos, net ir plačiajuosčio APWP atveju, išliko mažos ($0,05$ – $0,15$ %).



133 pav. Nuokrypio NC-RUS paklaidos mėginio storio įvertinimui esant 20 V sužadino lygiui (kairėje), 50 V sužadino lygiui (viduryje) ir 100 V sužadino lygiui (dešinėje)



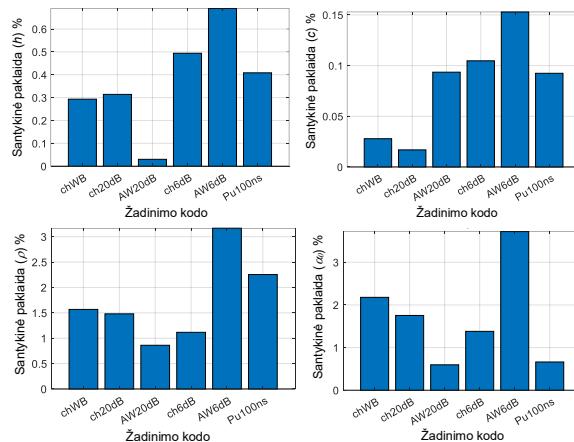
134 pav. Atsitiktinės NC-RUS paklaidos mėginio storio įvertinimui esant 20 V sužadino lygiui (kairėje), 50 V sužadino lygiui (viduryje) ir 100 V sužadino lygiui (dešinėje)

Šie rezultatai patvirtina, kad APWP su spektrine kompensacija yra tinkamiausias sužadino signalas NC-RUS matavimams visame analizuotame pralaidumo diapazone.

Panardinimo TRUS matavimų rezultatai

Panardinimo TRUS tyrimai patvirtino, kad APWP signalai su spektrinių nuostolių kompensavimu suteikia geriausią matavimo tikslumą esant vandens sujungčiai. Dėl labai mažos sužadino įtampos ir riboto pralaidumo panardinimo

sistemoje, APWP optimizavimas užtikrino aukštesnį SNR kritinėse dažnių srityse, palyginti su LFM ir NLFM signalais. Atlikus TRUS parametrų įvertinimą paaiškėjo, kad mažiausios paklaidos gaunamos naudojant plačiajuosį APWP su -20 dB kompensavimu, o siaurajuosčiai ir impulsiniai signalai sukelia daug didesnių sisteminių klaidų. Atskirti nuokrypio ir atsitiktinių paklaidų rezultatai, pateikti 18 pav., rodo, kad atsitiktinės paklaidos yra labai mažos ($0,025$ – $0,04$ %), o pagrindinį neapibrėžtumą lemia nuokrypio paklaidos, ypač amplitudės priklausomuose parametruose (tankis ir slopinimas).



135 pav. Nuokrypio TRUS paklaidos įvertinant mėginio storį (viršuje kairėje), greitį (viršuje dešinėje), tankį (apačioje kairėje) ir silpnėjimą (apačioje dešinėje)

Tai susiję su tuo, kad polikarbonato akustinis impedansas yra artimas vandens varžai, todėl rezonanso amplitudžių skirtumą įvertinti sudėtinga. Rezultatai aiškiai parodo nuostolių kompensuoto plačiajuosčio APWP signalo pranašumą panardinimo TRUS matavimuose.

Išvados

Buvo įrodyta, kad inversiniu filtru pagrįstas impulsų suspaudimas gali būti naudojamas oro jungiamuosiuose ir panardinamuosiuose TRUS matavimuose, siekiant išspręsti ilgų išsklaidytų spektro signalų persidengimo problemą.

Nuokrypio paklaidos gali būti sumažintos, jei naudojamas APWP signalas su nuostolių kompensavimu -20 dB pralaidumo juostoje. Oro jungties matavimuose storio įvertinimo nuokrypio paklaidos yra $0,18$ % plačiajuosčio APWP signalo ($0,38$ – $0,9$ MHz) atveju, palyginti su $0,45$ % to paties dažnio juostos pločio LFM signalo atveju ir $0,52$ % plačiajuosčio ($0,3$ – $1,1$ MHz) LFM signalo atveju, kai imties matavimui naudojamas 20 V sužadinimas. 100 V mėginio matavimuose APWP su nuostolių kompensavimu -20 dB nuostolių kompensavimu sukelia $0,03$ % paklaidą, palyginti su $0,45$ % tuo pačiu dažnių juostos pločio LFM signalu ir $0,62$ % plačiajuosčiu LFM. Atsitiktinės storio įvertinimo paklaidos yra nuo $0,05$ % iki $0,15$ % diapazone APWP su nuostolių kompensavimu -20 dB pralaidumo juostoje,

palyginti su 0,03 % iki 0,07 % diapazone plačiajuosčio LFM signalų. Tikėtina, kad APWP paklaidos bus didesnės, nes dalis energijos buvo panaudota spektriniais nuostoliams kompensuoti.

Panardinimo matavimuose storio įvertinimo paklaidos yra 0,04 % plačiajuosčio APWP signalo (0,74–3,43 MHz) atveju, palyginti su 0,31 % to paties dažnio juostos pločio LFM signalo atveju ir 0,3 % plačiajuosčio (0,3–3,9 MHz) LFM signalo atveju, esant 0,8 V sužadinimui. Atsitiktinės storio įvertinimo paklaidos yra 0,04 % APWP su nuostolių kompensavimu 20 dB pralaidumo juostoje, palyginti su 0,025 % plačiajuosčio LFM signalo atveju.

BENDROS IŠVADOS

1. Buvo sukurta apibendrinta APWP optimizavimo sistema, formuojanti sužadinimo spektrą taip, kad keitiklio pralaidumo juostoje būtų maksimaliai padidintas SNR. Palyginti su įprastiniais impulsiniais ir LFM signalais, APWP padidina naudojamą –20 dB pralaidumo juostą nuo maždaug 480 kHz iki 670 kHz, išlaikant tą pačią maksimalią įtampą. Taip pat buvo parodyta, kad spektrinei formai formuoti pakanka unipolinio sužadinimo, o tai reikšmingai supaprastina reikalingą elektroniką.

2. TRUS taikymuose SNR optimizuoti APWP signalai nuosekliai užtikrino mažiausias visų keturių įvertinamų parametrų (α , c , ρ , h) nuokrypio paklaidas. a) amplitudei jautrių parametrų (tankio, slopinimo) atveju APWP sumažino santykinės nuokrypio paklaidas 5–10 kartų, palyginti su LFM ir impulsiniu sužadinimu; b) laikui jautrių parametrų (greičio, storio) atveju paklaidos buvo sumažintos 3–15 kartų. Didžiausias pagerėjimas pasiektas oro sujungimo sąlygomis, kur vyrauja didelis slopinimas ir reverberacijos.

3. Spektrinės aprėpties analizė parodė slenkstinę poveikį nuokrypio paklaidoms, kai keitiklio pralaidumo juosta neapėmė mėginio rezonansinio piko ir gretimo slėnio, o įvertinti parametrai pasižymėjo didelėmis nuokrypio paklaidomis; juos įtraukus, nuokrypio paklaidos gerokai sumažėjo ir vėliau išliko beveik stabilios. Todėl tirtomis sąlygomis vienos piko-slėnio poros pakako stabiliam įvertinimui, o papildomų padengtų pikų įtraukimas reikšmingo tolesnio paklaidų sumažėjimo nesukėlė. Vertinant tik piką, gaunamos 7–9 % paklaidos. Įtraukus abu slėnius, gaunama mažiausia slopinimo paklaida (~3,8 %). Geriausias bendras kompromisas pasiekiamas tada, kai spektras apima visą –20 dB keitiklio pralaidumo juostą.

4. Buvo pasiūlytas ir patvirtintas impulsų suspaudimo metodas, pagrįstas atvirkštiniu filtru, skirtas persidengiančių atspindžių dekompozicijai ilgos trukmės išsklaidyto spektro signaluose. Kartu taikant impulsų suspaudimą, atidarymą ir dekompresiją, šis metodas sumažina nuokrypio paklaidas (greičio paklaidą nuo 0,085 % iki 0,002 %, o tankio paklaidą nuo 1,2 % iki 0,05 %, kai sužadinimo trukmė 350 μ s ir amplitudė 20 V).

5. Buvo sukurtas giluminio mokymosi modelis „ResidualUNet1D“, skirtas automatiniais atspindžiams atskirti ilgos trukmės išsklaidyto spektro signaluose. Naudojant 50 μ s trukmės signalus pasiektas ~45 dB likutinės energijos (RE) tikslumas

ir $\sim 0,9$ koreliacija, o naudojant 300 μ s trukmės signalus – ~ 32 dB RE ir $\sim 0,74$ koreliacija.

6. Panardinimo matavimuose storio įvertinimo nuokrypio paklaida yra 0,04 % plačiajuosčio APWP signalo (0,74–3,43 MHz) atveju, palyginti su 0,31 % to paties dažnių juostos pločio LFM signalo atveju ir 0,3 % plačiajuosčio (0,3–3,9 MHz) LFM signalo atveju, kai sužadavimo amplitudė yra 0,8 V. Atsitiktinės storio įvertinimo paklaidos yra 0,04 % APWP su nuostolių kompensavimu -20 dB pralaidumo juostoje, palyginti su 0,025 % plačiajuosčių LFM signalų atveju.

7. Siūloma APWP spektrinių nuostolių kompensavimo technika derinama su impulsų suspaudimu atvirkštinio filtro ir giluminio mokymosi pagrindu veikiančia dekompozicija, pagerina oro sujungimo TRUS matavimų kokybę ir tikslumą. Šie metodai leidžia patikimai išskirti rezonansinius parametrus naudojant tik vieną keitiklių porą plačiame dažnių diapazone, taip įveikiant ankstesnius apribojimus stipriai slopinančiomis oro sujungimo sąlygomis.

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CURRICULUM VITAE AND DESCRIPTION OF CREATIVE ACTIVITIES (CV)

Muhammad Tayyib

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Education:

2008 – 2010	Matriculation
2010 – 2012	Higher Secondary Education
2013 – 2017	Bachelors in Electronics Engineering
2017 – 2020	Masters in Electrical Engineering
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Professional experience:

2015 – 2016	Trainee Engineer at Creative Electronics and Automation, Islamabad, Pakistan.
2017 – 2020	Research Assistant at International Islamic University Islamabad, Pakistan.
2022 – 2023	Junior Researcher at Kaunas University of Technology, Lithuania.

Areas of research interest:

Nondestructive testing, resonant ultrasound spectroscopy, air-coupled ultrasound acquisition and processing, compressed sensing, plant physiology, time of flight measurements, optimization, adaptive digital signal processing.

Scientific papers related to the topic of dissertation:

1. **Muhammad Tayyib**, Linas Svilainis. Spectral content effects study in non-contact resonance ultrasound spectroscopy, *Sensors*, 2025, 25, 265. <https://doi.org/10.3390/s25010265> (Impact factor: 3.4, Q1).
2. **Muhammad Tayyib**, Linas Svilainis. SNR equalization in non-contact resonant ultrasound spectroscopy measurements. *NDT & E International*, 2025, 154. <https://doi.org/10.1016/j.ndteint.2025.103386> (Impact factor: 4.1, Q1).
3. Linas Svilainis, Muhammad Tayyib, Pulse compression for long spread spectrum signals application in air-coupled ultrasonic thickness resonance spectroscopy, *IEEE Transactions on Instrumentation and Measurement*, 2026, <https://doi.org/10.1109/TIM.2026.3676189> (Impact factor: 5.9, Q1).

Scientific conferences:

1. **Muhammad Tayyib**, Linas Svilainis, Signal to noise ratio equalization in Ultrasonic air-coupled resonance spectroscopy by programable spectral shape. ULTRASONICS 2023 6th International Caparica Conference on Ultrasonics 26th – 29th June 2023, Caparica, Portugal
2. **Muhammad Tayyib**, Linas Svilainis, Plant physiological parameter estimation using resonant ultrasound spectroscopy with improved bandwidth and

enhanced SNR. 17th Annual International Symposium on Agricultural Research 24-28 June 2024, Athens Greece.

3. **Muhammad Tayyib**, Linas Svilainis, Compensation function for enhanced bandwidth and improved SNR through programmable spectral shape signals. IEEE Ultrasonics, Ferroelectrics and Frequency Control Joint Symposium 22-26 September 2024, Taipei, Taiwan. DOI: 10.1109/UFFC-JS60046.2024.10793903

4. **Muhammad Tayyib**, Mussayab Yousufi, Linas Svilainis, Andrius Chaziachmetovas Enhancing air-coupled ultrasonic inspection: resolving signal overlapping in long-duration chirp signals, 2025 International Congress on Ultrasonics, 21-25 September 2025, Paderborn, Germany. DOI: 10.5162/Ultrasonic2025/A16-a4

LIST OF SCIENTIFIC PAPERS AND SCIENTIFIC CONFERENCES

Articles indexed in the Web of Science with impact factors

1. TAYYIB, M.; SVILAINIS, L. Spectral content effects study in non-contact resonance ultrasound spectroscopy. *Sensors*, 2025, 25, 265. <https://doi.org/10.3390/s25010265>
2. TAYYIB, M.; SVILAINIS, L. SNR equalization in non-contact resonant ultrasound spectroscopy measurements. *NDT & E International*, 2025, 154. <https://doi.org/10.1016/j.ndteint.2025.103386>
3. SVILAINIS, L.; TAYYIB, M. Pulse compression for long spread spectrum signals application in air-coupled ultrasonic thickness resonance spectroscopy. *IEEE Transactions on Instrumentation and Measurement*, 2026. <https://doi.org/10.1109/TIM.2026.3676189>

List of conference presentation

1. TAYYIB, M.; SVILAINIS, L. Signal to noise ratio equalization in ultrasonic air-coupled resonance spectroscopy by programmable spectral shape. In: 6th International Caparica Conference on Ultrasonics, 26–29 June 2023, Caparica, Portugal.
2. TAYYIB, M.; SVILAINIS, L. Plant physiological parameter estimation using resonant ultrasound spectroscopy with improved bandwidth and enhanced SNR. In: 17th Annual International Symposium on Agricultural Research, 24–28 June 2024, Athens, Greece.
3. TAYYIB, M.; SVILAINIS, L. Compensation function for enhanced bandwidth and improved SNR through programmable spectral shape signals. In: IEEE Ultrasonics, Ferroelectrics and Frequency Control Joint Symposium, 22–26 September 2024, Taipei, Taiwan. <https://doi.org/10.1109/UFFC-JS60046.2024.10793903>
4. TAYYIB, M.; SVILAINIS, L. Smart farming redefined: Portable air-coupled ultrasonic technology for precision agriculture. In: 21st International Conference of Young Scientists on Energy and Natural Sciences Issues (CYSENI), 20–22 May 2025, Kaunas, Lithuania.
5. TAYYIB, M.; YOUSUFI, M.; SVILAINIS, L.; CHAZIACHMETOVAS, A. Enhancing air-coupled ultrasonic inspection: Resolving signal overlapping in long-duration chirp signals. In: 2025 International Congress on Ultrasonics, 21–25 September 2025, Paderborn, Germany. <https://doi.org/10.5162/Ultrasonic2025/A16-a4>
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for ultrasound spectroscopy. In: 2025 International Congress on Ultrasonics, 21–25 September 2025, Paderborn, Germany. <https://doi.org/10.5162/Ultrasonic2025/C3-a1>

8. TAYYIB, M.; YOUSUFI, M.; SVILAINIS, L. Non-invasive monitoring of plant water status using portable air-coupled ultrasonics for precision agriculture. In: 14th International Conference of Young Scientists, 26 November 2025, Vilnius, Lithuania.

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