Piezoelectric transceiver matching for multiple frequencies

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Abstract. Robust data transmission over long ranges with standard ultrasound devices is a challenge. Ultrasound indoor positioning systems in particular require long ranges and a robust data communication link. Fundamentally, a piezoelectric transducer has a narrow bandwidth for high sound pressure level and efficiency and is not suitable for broad-band applications. Moreover, ultrasound attenuation in the air increases quadratically within frequency, and thus ultrasound localization systems are restricted to low frequencies and low bandwidths.

This work presents a novel method to match a piezoelectric transceiver for multiple frequencies by using the parallel and the series resonance of the transceiver. The aim is to adjust the amplitudes at different frequencies from different senders to the same level, which is important for orthogonal frequency division multiplex communication systems. Hence, an analog-to-digital converter (ADC) with low dynamic range (low voltage resolution) can be used to measure multiple frequencies with the same resolution. As a result, the optimization decreases the required dynamic range by 6 dB. Consequently, the ADC requires 1 bit fewer to ensure the same resolution for all carrier frequencies.

1 Introduction

Ultrasound seems to be a preferable technology for indoor localization. It is low cost, the speed of sound is about six decades slower than the speed of light and it is limited to closed rooms. Lerch et al. (2008) and Tränkler and Reindl (2014) presented systems for distance measurements and object recognition. Hence, high localization accuracy can be achieved with low complexity devices and a low sampling frequency analog-to-digital converter (ADC) due to the low velocity of sound propagation. Furthermore, the sound distribution is limited to small areas (e.g., rooms). However, the expected maximal distance for ultrasound is about 20 m and the bandwidth is small compared to radio frequency-based systems.

Piezoelectric transducers have a high quality factor $Q$ and high sound pressure levels (SPL) up to 120 dB (in 0.2 m distance at only $U_{pp} = 20$ V), whereas the usable bandwidth is small and enables only narrow bandwidth application. Hence, Ens et al. (2013) developed a modulation scheme that is based on frequency shift keying (FSK) and chirp spread spectrum (CSS) techniques to achieve robust data transmission with narrow bandwidth devices. However, Jakes (1993) showed that multi-carrier data communication provides higher spectral efficiency than CSS modulation. Therefore, we are interested in adapting the system for multi-carrier data communication in an orthogonal frequency division multiplex (OFDM). Chang (1966) and Weinstein and Ebert (1971) showed the possibilities and advantages of OFDM. Transmission systems require transducers with higher bandwidth or the use of multiple transducers with different resonance frequencies. Hosman et al. (2011) showed the use of a standard 40 kHz piezoelectric ultrasound transducer for data communication in metal within a frequency band of 260 to 330 kHz. Providing such high frequencies to ultrasound, the signal (200 kHz) will be attenuated by about 60 dB (about 33 dB for 40 kHz) at a 5 m distance (ISO, 1993). However, we use lower frequencies (40 kHz) for our data transmission to reduce the attenuation by air.

Oralkan et al. (2002) and Caliano et al. (2010) investigated the bandwidth of capacitive micro-machined ultrasound transducers (CMUT) and showed the broad-band capabilities. They have a wide-band characteristic and can be
fabricated with standard silicon technology. Furthermore, the CMUTs are designed to operate below the resonance frequency of the device. This enables higher bandwidths. CMUTs include a low noise amplifier (LNA) to gain the signal. Hence, the device requires no matching and the signal can be amplified with an operational amplifier. However, the noise floor of the CMUT devices is higher than for piezoelectric devices. Furthermore, Anderson et al. (2005) showed that the CMUTs require high voltages to transmit signals and they have a low efficiency.

Bloomfield et al. (2000) presented ultrasound transducers based on polyvinylidene fluoride (PVDF) with high bandwidth and a relatively good acoustic impedance match to water. However, they only have an output of about 0.025 Pa V−1 and require a supply voltage of about 200 Vpp to achieve a SPL of 107 dB; one example is the US40KT-01 transducer by MSI.

Ealo et al. (2008) showed the use of a microporous polypropylene foam with permanent charge and a broadband characteristic as an electromechanical film (EMFi). The transducer is customizable and consists of a film glued to a rigid surface and the connections for an external supply voltage. However, there are no commercially available EMFi-transducers and the high supply voltage makes it unsuitable for low-power application.

Manufacturers provide ultrasonic transducer in pairs for sending and receiving. They design the device so that the sender has a series resonance near the same frequency as the parallel resonance of the receiver. The presented work shows how to use both transducers for sending on two frequencies and only one transducer for receiving the signal. Moreover, it shows an optimization method with a measurement resistor to use for a series and a parallel resonance circuit for multi-frequency applications. This includes the calculation of the measurement resistor parallel to the piezoelectric transducer. The transmission system receives two signals with two different frequencies, which are also generated with piezoelectric transducers. The aim of the optimization is to match both signals to the same amplitude. Hence, a low-cost ADC with lower dynamic range can be used to convert the signals to digital values.

2 Model

The system is designed to transmit data over two separate channels with different carrier frequencies in OFDM. Figure 1 shows the schematic diagram of the transmission system. The data bits \( b_0 \) and \( b_1 \) are mapped by \( \pi/4 \)-BPSK (binary shift keying; Höher, 2013), and the pulse width modulation (PWM) generates the analog signal. Additionally, the power amplifiers (PA) gain the signal before the piezoelectric transducers transform the electric signal to ultrasound signal. The composite signal traveling within air medium is transformed to electric signals at the receiver by the piezoelectric transducer. In addition, the air is assumed to be linear and the attenuation difference between the neighboring frequencies is negligible due to the small frequency difference, the low amplitudes (below 120 dB SPL) and the interest in relative amplitudes. Moreover, the additive white Gaussian noise (AWGN) \( w_n \) adds the noise to the received signal. The low noise amplifier (LNA) with bandpass filter (BF) gains the received signal by a factor of 18. The analog-to-digital converter (ADC) generates the digital data from the analog signal and the demapper estimates the data bits.

The equivalent circuit for the piezoelectric transceiver is based on the Butterworth–van Dyke model (Van Dyke, 1928; Mason, 1935) see also the IRE standard (IRE, 1957). Figure 2 shows the equivalent circuit of the piezoelectric senders and the receiver. Thus, the piezoelectric transducer model includes a resistor, an inductance and two capacitances. In this work both transducers are used, the receiver (MA40S4R) and the sender (MA40S4S), for generating the ultrasound signal. However, only one transducer is used to receive the superposition of the signals from both devices.

2.1 Sender

The sender and receiver have the same equivalent circuit, for which each has different values. Therefore, the components are indexed by \( kx \) with \( k \in \{1, 2\} \), whereby \( k = 1 \) corresponds to the sender device and \( k = 2 \) to the receiver device. The impedance of the sender is derived from the elements in series.
The transmitted acoustic power of the sender is proportional to the dissipated power of the resistor $R_{k_1}$. Hence, the transfer function of the energy depends on the values of the capacitance $C_{k_1}$, the inductance $L_{k_1}$ and the resistor $R_{k_1}$:

$$Z_{T_s} (\omega, k) = \frac{R_{k_1} + \frac{1}{j \omega C_{k_1}} + j \omega L_{k_1}}{1 + \frac{C_{k_2}}{C_{k_1}} + j \omega R_{k_1} C_{k_2} - \omega^2 C_{k_2} L_{k_1}}.$$  

(3)

The measurement resistor $R_{m}$ is connected in parallel to the piezoelectric transceiver (MA40S4R). The series and parallel resonance frequencies are independent of the measurement resistor. However, the measurement resistor influences the amplitude of the received signals. Hence, the usable carrier frequencies should be in the region of the resonance frequencies to receive signals with low loss and high amplitudes. The series impedance depends on the angular frequency $\omega = 2 \pi f$ and includes the resistor $R_{21}$, series inductance $L_{21}$ and series capacitance $C_{21}$:

$$Z_{R_1,s} (\omega) = R_{21} + \frac{1}{j \omega C_{21}} + j \omega L_{21}.$$  

(6)

Furthermore, the voltage source in series with the resistor $R_{21}$, the capacitance $C_{21}$ and the inductance $L_{21}$ represents the transformation of the acoustic signal into an electric signal. The sink for the signal is the parallel impedance which includes the measurement resistor $R_{m}$ and the parallel capacitance $C_{22}$:

$$Z_{R_1,p} (R_{m}, \omega) = \frac{1}{\frac{1}{R_{m}} + j \omega C_{22}} = \frac{R_{m}}{1 + j \omega C_{22} R_{m}}.$$  

(7)

Moreover, the transfer function from the source of the piezoelectric generator $U_{R_s}$ to the measurement resistor is

$$G_{R_s} (R_m, \omega) = \frac{Z_{R_1,p} (R_m, \omega)}{Z_{R_1,s} (\omega) + Z_{R_1,p} (R_m, \omega)}.$$  

(8)

2.3 Parameter characterization

For a numerical simulation of the presented model, the parameters of the piezoelectric transducers are calculated and characterized. Thus, the impedance of the receiver (Murata MA40S4R) and sender (Murata MA40S4S) devices is measured. Figure 3 shows the measured magnitudes of the impedances of both devices. The minimum of the magnitude corresponds to the series resonance and the maximum of the magnitude corresponds to the parallel resonance. Figure 4 shows the phase angle of the impedance of the piezoelectric sender and receiver. The phase crosses the abscissa at the resonance frequencies. Therefore, the imaginary part at the resonance frequencies is zero.

The values of $C_{21}$ and $L_{21}$ can be calculated approximately with the series resonance frequency $f_s$ of the series resonance circuit (IRE, 1957)

$$f_s = \frac{1}{2 \pi \sqrt{L_{21} C_{21}}}.$$  

(9)
Figure 3. Impedance magnitude of the piezoelectric receiver and sender. The minimum magnitude corresponds to the series resonance and the maximum to the parallel resonance. The series resonance of the sender is near the parallel resonance of the receiver.

Figure 4. Impedance phase angle of the piezoelectric receiver and sender. The zero-crossings indicate the resonance frequencies.

The capacitance $C_{22}$ of the parallel capacitance can be calculated with Eqs. (9) and (10) as

$$C_{21} = C_{22} \left( \frac{f_p^2}{f_s^2} - 1 \right).$$

Furthermore, the value of the inductance with the series resonance frequency $f_s$ and the capacitance $C_{21}$ is calculated with Eq. (11) as

$$L_{21} = \frac{1}{\omega_s^2 C_{21}}.$$

The quality factor $Q$ describes the damping of the resonance circuit depending on the capacitance and the resistor in the series resonance or on the resonance frequency and the 3 dB bandwidth (IRE, 1957):

$$Q = \frac{1}{\omega_s C_{21} R_{21}} = \frac{f_s}{b}.$$

### Table 1. Measured and calculated values of the transceiver Murata MA40S4S and MA40S4R.

| Sender MA40S4S, $k = 1$ | Receiver MA40S4R, $k = 2$ |
|--------------------------|---------------------------|
| $f_s$                    | 40.8 kHz                  | 38.8 kHz                  |
| $f_p$                    | 43.5 kHz                  | 41.5 kHz                  |
| $R_{k1}$                 | 204 $\Omega$              | 240 $\Omega$              |
| $L_{k1}$                 | 43.6 mH                   | 45.8 mH                   |
| $C_{k1}$                 | 349 pF                    | 367 pF                    |
| $C_{k2}$                 | 2.55 nF                   | 2.55 nF                   |
| $Q$                      | 54.8                      | 46.6                      |

Table 1 shows the measured and calculated values for the receiver and sender transducers.

### 3 Simulation

The aim of the simulation is to find the optimal value of the measurement resistor $R_m$ to receive both signals on two carrier frequencies ($f_0 = 38.8$ kHz and $f_1 = 40.8$ kHz) with the same signal strength. The received signal strength for every frequency depends on the value of the measurement resistor. In detail, a high value of the measurement resistor causes high amplitudes for signals received within the parallel resonance circuit. Indeed, a low value of the measurement resistor damps the parallel resonance circuit and the current in the series resonance circuit flows. Furthermore, the signal strength at the senders differs for the frequencies. Hence, the value has to be optimized regarding the ratio of the amplitudes $r = \frac{A_{0m}}{A_{0n}}$. $A_{0n}$ and $A_{0m}$ are the amplitudes of the signals at the implemented frequencies, which can be determined with a broad-band microphone. Moreover, the received signal $U_{R_k}$ consists of the transmitted voltage of both senders:

$$U_{R_k} = U_{R_{k1}} + U_{R_{k2}}.$$  \hspace{1cm} (14)

Hence, the transfer function from the sender to the receiver is

$$G_m(R_m, k) = G_{T_k}(R_m) \cdot G_{R_k}(k).$$  \hspace{1cm} (15)

Therefore, the condition for optimal signal strength, with respect to the used frequencies $f_0$ and $f_1$, is

$$\text{arg min}_{R_m} | Re \{G_m(R_m, 1) \cdot r - Re \{G_m(R_m, 2)\} |.$$  \hspace{1cm} (16)

Thus, the optimal resistor follows through minimization:

$$R_{m, \text{opt}} = \arg \min_{R_m} | Re \{G_m(R_m, 1) \cdot r - Re \{G_m(R_m, 2)\} |.$$  \hspace{1cm} (17)

Figure 5 shows the numerical simulation results of the received voltage at the measurement resistor $R_m$ for both frequencies ($f_0$ and $f_1$) at different values of the measurement resistor. The curve of the voltage at the frequency $f_1$, which
is near the parallel resonance frequency, has a higher slope than the voltage of the frequency $f_0$. Furthermore, the voltage of the frequency $f_0$ (near to the series resonance) saturates to 0.9. As a result, the optimal value of the measurement resistor is $R_m \approx 1452 \Omega$, the intersection of both curves.

Moreover, a simulation of the transfer function for the optimal measurement resistor value was also performed. Figure 6 shows the simulation results for unmatched ($R_m = 1 \times 10^6 \Omega$) and matched ($R_m = 1452 \Omega$) frequency response of the piezoelectric transmission system. An unoptimized system with $R_m = 1 \, \text{M} \Omega$ requires a dynamic range 10 dB higher than an optimized system with $R_m = 1.452 \, \text{k} \Omega$ to measure both signals with same resolution. Therefore, the ADC requires 10 dB less dynamic range to measure both signals. Figure 7 shows the phase of the transfer function for the matched and unmatched receiver. The phase angle for $f_0 = 38.8 \, \text{kHz}$ remains constant at about $0^\circ$ for both situations. However, the phase angle for $f_1 = 40.8 \, \text{kHz}$ changes from about $-50^\circ$ for the unmatched case to $-80^\circ$ for the matched receiver. At the series resonance of $f_0 = 38.8 \, \text{kHz}$ the imaginary part of the receiver device (MA40S4R) is about zero and the value of the resistor $R_{21}$ is dominant. Hence, the influence of the measurement resistor is marginal at the series resonance. Though, at the parallel resonance at $f_1 = 40.8 \, \text{kHz}$ of the receiver device (MA40S4R), the imaginary part increases to the maximum value and the measurement resistor $R_m$ in parallel becomes dominant. In addition, the measurement resistor $R_m$ influences the phase and the amplitude of the signal at the frequency $f_1 = 40.8 \, \text{kHz}$. As a result, the reduction of the measurement resistor $R_m$ also reduces the real part and causes the phase shift of about $30^\circ$.

### 4 Measurements

The impedance and the transfer function of the system are measured to verify the simulation results and the capability of transmitting data. Therefore, the senders are placed together to act as a point source in line with the receiver within a distance of 30 cm.
The impedance of the receiver is measured to show the changes by the measurement resistor. Figure 10 shows the magnitude and Fig. 11 the phase angle of the impedance for the selected measurement resistor \( R_m = 1 \, \text{k}\Omega \), \( R_m = 1.4 \, \text{k}\Omega \) and \( R_m = 1 \, \text{M}\Omega \) (black curve). When increasing the measurement resistor, the phase angle is more flattened. Furthermore, the magnitude of the impedance of the parallel resonance circuit decreases to the value of the measurement resistor. The phase at the series resonance has a higher slope than at the parallel resonance. Therefore, the parallel resonance can be better adjusted to the used frequency. Hence, the low frequency signal \( f_0 \) has to be near the series resonance frequency, whereas the high frequency signal \( f_1 \) can be in the range of about 1 kHz to the parallel resonance frequency of the piezoelectric transceiver.

The transmission system in Fig. 1 is matched by the optimal resistor with 1.4 k\Omega. The senders transmit two data bits \((b_0 = 0 \text{ and } b_1 = 1)\) with the symbol duration of 1 ms and with the amplitude of 8 Vpp over 1.3 m. Hence, the phase shift is shifted for the carrier frequency \( f_0 \) by \( \pi + \pi/4 \approx -2.36 \) and for \( f_1 \) by \( \pi/4 \approx 0.79 \). The first symbol is used as the reference symbol to calculate the phase angle difference at the receiver after the ADC. Before the first symbol there is the eigen oscillation of the receiver with the piezoelectric transducer and the bandpass filter. In the first symbol, the difference frequency of 2 kHz can be recognized by the periodic change of the envelope with the period of 0.5 ms. Moreover, the amplitude at 0.8 ms decreases nearly to zero, which indicates that the amplitudes of the carrier frequencies are similar. At the beginning of the second symbol, the phase angles change at both carrier frequencies and the amplitude decreases. After the first half of the symbol duration the phases angles are stabilized again and the envelope indicates the 2 kHz difference frequency. As a result, the matching provides an optimal signal for data transmission with low complexity and effort. At the end of the transmission the senders and the receiver swing off, which is indicated as ringing.

Figure 13 shows the phase angle of the signal over time. The phase angle remains stable after the half of the first symbol. Thus, the phase is normalized to 0° at the center of the first symbol. The phase angle varies at the begin-
with a measurement resistor of $R_m = 1300 \, \Omega$ for a signal with two different frequencies (38.8 and 40.8 kHz) shows the same amplitude. Moreover, the signals have the same amplitude and the same group delay (1.36 ms) for the matched transducer. As a result, the optimization reduces the required dynamic range of the ADC by 6 dB. Hence, the ADC requires 1 bit fewer to ensure the same resolution for both carrier frequencies. Furthermore, the measurements validate our model and the optimization with numerical simulation. Data transmission works over two carrier frequencies with two senders and one receiver. Consequently, transmission systems based on multiple carrier frequencies may use piezoelectric transducers wherein the series resonance frequencies of the senders device lie in the range of the series and the parallel resonance frequency of the receiver device.

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