Published in final edited form as:

Phys Med Biol. 2008 July 21; 53(14): 3943–3969. doi:10.1088/0031-9155/53/14/014.

Efficient Array Design for Sonotherapy

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Abstract

New linear multi-row, multi-frequency arrays have been designed, constructed and tested as fully operational ultrasound probes to produce confocal imaging and therapeutic acoustic intensities with a standard commercial ultrasound imaging system. The triple-array probes and imaging system produce high quality B-mode images with a center row imaging array at 5.3 MHz, and sufficient acoustic power with dual therapeutic arrays to produce mild hyperthermia at 1.54 MHz. The therapeutic array pair in the first probe design (termed G3) utilizes a high bandwidth and peak pressure, suitable for mechanical therapies. The second multi-array design (termed G4) has a redesigned therapeutic array pair which is optimized for high time-averaged power output suitable for mild hyperthermia applications. The “thermal therapy” design produces more than 4 Watts of acoustic power from the low frequency arrays with only a 10.5 °C internal rise in temperature after 100 seconds of continuous use with an unmodified conventional imaging system, or substantially longer operation at lower acoustic power.

The low frequency arrays in both probe designs were examined and contrasted for real power transfer efficiency with a KLM model which includes all lossy contributions in the power delivery path from system transmitters to tissue load. Laboratory verification was successfully performed for the KLM derived estimates of transducer parallel model acoustic resistance and dissipation resistance, which are the critical design factors for acoustic power output and undesired internal heating respectively.

1. Introduction

Ultrasound imaging combined with ultrasonic therapy has the potential to treat a wide range of diseases, including cancer, peripheral vascular disease, and stroke. (Mitravotri, 2005; Pitt *et al.*, 2004) While high temperatures are required to ablate tissues, mild hyperthermia can be used to change the properties of a drug delivery vehicle and increase local perfusion. (Jones *et al.*, 2006; Ferrara, 2008)

In order to maintain the low cost, portability, and real-time nature of ultrasound, a high quality ultrasound image should be obtained within the target region, guiding the mechanical or thermal therapeutic pulses. Since many regions within the body allow only a small acoustic window for both imaging and therapeutic access, the development of confocal imaging and therapeutic arrays is highly desirable. Also, given the sophistication and complexity of current ultrasound imaging systems (facilitating many modes of tissue and blood flow imaging), the incorporation of therapeutic arrays within a high quality imaging system remains an important goal.

Sonotherapy itself is a general term describing the use of sound as a direct medical treatment, or as a mediator in a treatment. The use of ultrasound for therapy is not new, but ultrasound applications in innovative drug delivery methodologies is a promising area of research to enhance the effectiveness of, for example, cancer drugs, at a specific site of a known or suspected tumor (Ferrara, 2008; Jones *et al.*, 2006; Pitt *et al.*, 2004). Ultrasound mediation of drug delivery can be exploited in several ways; our lab has explored two such methods as briefly described here.

The transmission of ultrasound pulses with a low center frequency, short pulse length and high peak pressure, results in a local mechanical stress and can be employed for mechanically-optimized therapies, such as those using ultrasound contrast agents. Probe designs described in the last several years (Stephens *et al.*, 2006; Zheng *et al.*, 2007; Zhao *et al.*, 2007; Stieger *et al.*, 2007) have provided both visualization and contrast agent mediated drug delivery.

Alternatively, the transmission of pulses with a long pulse length and low peak pressure can be used in thermal therapies, in which local heating changes perfusion, enhances angiogenesis, or releases a drug from thermally-activated drug delivery vehicles. Although heating increases with increasing ultrasound center frequency, the use of an ultrasound center frequency in the low megahertz range is frequently proposed for tissue heating in order to avoid overheating the skin or other sensitive tissues and to facilitate treating larger regions of interest within a reasonable time frame. One form of thermally-activated carriers, now in pre-clinical evaluation is the liposome. Liposomal encapsulation of a drug can act as a reasonably stable, biocompatible drug delivery carrier with lower systemic toxicity and better drug containment compared with free drug (Allen, 1997). The drug carrying liposomes can be easily injected into a patient's blood stream and are small enough to extravasate in leaky tumor vasculature. With a long circulating temperature sensitive liposome (TSL), the vehicles can deliver their drug payload when a targeted region of the body is elevated in temperature by a few degrees (Dromi *et al.*, 2007). The hyperthermia used to release a drug from a thermally-activated particle need not be maintained for an extended period—if the particles have accumulated at the target site heating can be completed in a few minutes. Therefore, a practical therapeutic regimen requiring a short interval of heating can be proposed.

Several technical challenges have emerged with this requirement for greater time-averaged power. First, a reasonable operating bandwidth is required in the low frequency arrays, along with a high output power design with low internal probe heating. The design should also work well with a commercial imaging system without the need for encumbering system modifications. Generally speaking, array designs for high power use have been optimized for very narrow band operation, but with only minor attention to the minimization of internal heating. The “power arrays” in the designs considered here are positioned in the probe head within 1 mm on either side of the center imaging array. This limitation alone makes the design very challenging with regard to excessive heat management.

This work considers two design generations, a design optimized for mechanical therapy (the G3 design) in which the outer rows are made of a triple piezoceramic layer construction with multiple matching layers for wide bandwidth and high peak pressure; and a new G4 thermal design with a standard single layer piezoceramic construction with a single high thermal conductivity matching layer.

The study here addresses key issues in the design of sonotherapy delivery devices, but also has fundamental application to any acoustic transducer design. The principal goal of this study is to validate both our detailed KLM acousto-electric model and the design of the newer (G4) device which is constructed for high efficiency sonotherapy. A sonotherapy device should deliver a satisfactory acoustic output for its application with high efficiency and minimal array

heating. In particular, the transducer acoustic radiation resistance, R_a , the transducer dissipation resistance, R_d , and the system source resistance, R_s , are considered in this study to address:

- a. the means to calculate and measure both components of the classical transducer motional resistance, R_m , where R_m is the parallel sum of R_a , and R_d , and
- b. the effects of R_s , R_a , and R_d on array operation efficiency in two distinctly different high quality array designs.

Using frequency domain quantities, it will be shown that for a given design, the ratio of R_d and R_a weighted with the acoustic power output can be used to demonstrate a spectral peak which indicates a maximum of real power output and a minimum of undesired internal heating.

2. Methods

2.1 Array Design for Targeted Drug Delivery

The general design format of the multi-frequency, multi-row array has been described previously (Stephens *et al.*, 2006) as a novel probe design which offers high quality imaging at 5.3 MHz with the central array, and power delivery with a pair of 1.5 MHz arrays on either side of the center imaging array. A picture of the “CoLinear” probe face is shown in Figure 1a, with all three individual array positions outlined. Figure 1b shows the center row array imaging capability during a therapeutic study; the center row array images while the outer rows are used in spectral Doppler mode for heating a specific small volume of tissue; the two modes are utilized in rapid alternating fashion in much the same way that combined imaging and Doppler modes works on a standard imaging system.

The inner array is comprised of 128 elements in a single piezoceramic layer designed to operate at 5.3MHz for high quality imaging. The two low frequency outer rows, one on each side of the center row, are hard-wired together in parallel with each pair of 64 corresponding elements. This array configuration was chosen because of its reasonably straight forward adaptation from a single frequency three row 1.5D connectivity array to our multifrequency arrangement. No elevational lensing is used; the flat polyurethane protection layer has negligible effects on the elevation beam.

The inner and outer array acoustic beams are designed to overlap in the elevation plane at 3.5 cm (for breast cancer applications) from the probe face with the outer arrays mechanically oriented towards the center axis with a design angle of 10.2 degrees. Neither the inner or outer arrays have active elevational lensing on the probe front face, the polyurethane encapsulation is impedance matched to body tissue with low attenuation.

Both the G3 and G4 have been fully modeled, constructed, and laboratory tested. The probes are equipped with a standard ZIF connector housing with tuning inductor boards and standard manufacturing cabling. The Sonoline Antares™ (Siemens Medical Solutions, USA) was used as the interface system with minor modifications to system software to enable the separate, but coordinated use of the inner and outer arrays in both the G3 and G4. In addition to the software modifications, the Antares Diagnostic User Interface (DUI) was used to obtain direct control over various parameters such as voltage (V_{DUI}), PRF, number of transmit cycles, aperture control, etc.

2.2 Array Design Requirement and Modeling for Efficiency

A solution to the Pennes bio-heat equation (Nyborg, 1988) (Kruse *et al.*, 2007) was used to compute a reasonable estimate of the acoustic output power needed to heat a small volume (i.e. about 0.5 cm by 0.5 cm beam area) of superficial tissue. Considering conservatively high estimates for blood perfusion, it was determined that 4 Watts of array-generated focused power

delivered to our small volume of interest would satisfactorily elevate the tissue temperature by approximately 7 °C in 100 seconds of insonation.

The power performance optimization of the outside array pair is the principal subject of this study; references to an array from this point are intended to specify the low frequency array pair in either the G3 or G4 probe designs.

A “system based” KLM model from prior work (Lamberti *et al.*, 1997; Chopra *et al.*, 2003; Selfridge and Gehlbach, 1985; Oakley, 1997; Lockwood and Foster, 1994; Wu and Chen, 1999; Chen, 2002) was used to include the Antares system transmitters, cable, array, and acoustic and electrical loading. This model has produced good agreement between laboratory observations and model results.

2.3 Complete Model with System Parameters

Since it is the acoustic power delivery efficiency capability of the low frequency arrays which is of principal concern, the major sources of power loss along the transmission pathway have been incorporated into the KLM model in both the G3 and G4 designs.

To accomplish tissue heating, the two 64 element arrays work essentially in parallel, each element of one array is hard-wired to the other array's corresponding element. For modeling purposes, each dual-array element exists in parallel, but also all 64 pairs are in (phased) parallel operation with 64 transmitters and 64 inductors. For purposes of 1-D modeling, the entire 2 outer arrays are considered as a “single element” in operation which is reasonable since the outer arrays together are used in phase unison to achieve the maximum benefit from the production of acoustic power.

The loss components in the power transmission path to the probe are the system transmitter series output resistance, R_s , the tuning inductor resistance which is estimated as

$$R_L = 2\pi f_i \cdot L / Q_L, \quad (1)$$

and the cable resistance, R_{cab} , which appears in both the cable characteristic impedance and cable propagation constant. The transducer itself has losses in every layer considered, which are characterized by the mechanical Q_m defined for each layer, and the backing material. A particular layer's Q_m can be used to derive an estimated real term in the acoustic propagation constant at a relevant characterization frequency, f_m , by

$$\alpha(f_m) = \frac{4.343 \cdot 2\pi f_m}{Q_m \cdot \text{vel}_m \cdot 8.69} \quad (2)$$

and in the derivation of the complex acoustic impedance for the KLM model as

$$Z_{KLM} = Z_{\text{layer}} \cdot A_{xd} \cdot \left(1 - \frac{j}{2 \cdot Q_m}\right). \quad (3)$$

The electrical loss in the piezoceramic is characterized by the loss tangent for the material used, and appears in the dielectric constant as

$$\epsilon = \epsilon_0 \epsilon_r (1 - j \tan \delta). \quad (4)$$

Figure 3 shows the outline of the 2X2 matrix elements used in the model of a single transducer element pair driven with its system components. The model for the entire network can be

considered as three general matrix elements: the system series network, $M1$, which includes the Antares system output impedance, tuning inductor and cable characteristics, the transducer matrix, $M2$, with its backing and matching layers, and the series element load matrix $M3$ which is the medium into which the acoustic power is delivered.

For the purposes of examination of the effects of the system transmitter, tuning inductor and the cable, the KLM matrices of the source network are defined as

$$M1a(f) = \begin{bmatrix} 1 & R_s \\ 0 & 1 \end{bmatrix}, \quad (5)$$

$$M1b(f) = \begin{bmatrix} 1 & Z_L \\ 0 & 1 \end{bmatrix}, \quad (6)$$

$$M1c(f) = \begin{bmatrix} \cosh(\gamma_{cab}(f)) & Z_{cab}(f) \cdot \sinh(\gamma_{cab}(f)) \\ \frac{1}{Z_{cab}(f)} \sinh(\gamma_{cab}(f)) & \cosh(\gamma_{cab}(f)) \end{bmatrix}, \quad (7)$$

where R_s is the output resistance of the system transmitters, Z_L is the complex impedance of the series tuning element in the probe's system-end connector, $\gamma_{cab}(f)$ is the complex frequency dependent propagation constant for the cable length, and $Z_{cab}(f)$ is the frequency dependent cable characteristic impedance. Refer to Tables 3 through 6 for design values used for the G3 and G4 array designs.

The transducer network itself is summarized as the product of the well known 2X2 matrix components of the KLM model for a piezoceramic transducer stack with backing and matching layer included, which is defined here and in Figure 3 to be

$$MX(f) = M2a(f) \cdot M2b(f) \cdot M2c(f) \cdot M2d(f) \cdot M2e(f) \cdot M2f(f) \cdot M2g(f). \quad (8)$$

The summary matrix for the whole system, $M_{total}(f)$, can be arranged with the tissue impedance as the final series load, as

$$M_{total}(f) = M1a(f) \cdot M1b(f) \cdot M1c(f) \cdot MX(f) \cdot \begin{bmatrix} 1 & Z_{tissue} \\ 0 & 1 \end{bmatrix}, \quad (9)$$

and additionally a useful relationship can be arranged here for use later in the calculation of the loaded cable input impedance, as

$$MTCX(f) = M1b(f) \cdot M1c(f) \cdot MX(f). \quad (10)$$

An overall system transmit-ratio spectrum function can now be found by taking the “1/B” term in the resulting 2X2 matrix, which is interpreted as the particle velocity out of the transducer divided by the transmit voltage from the system pulsers, or

$$Tx_ratio(f) = 1 / [M_{total}(f)_{0,1}] \quad (11)$$

Using this relationship, the transmission transfer function (TTF) can be found as the force on the load produced by the array for the transmission drive voltage, or

$$TTF(f) = \text{Force}_{tissue} / V_{DUI} = Tx_ratio(f) \cdot Z_{tissue} \quad (12)$$

where V_{DUI} is the transmitter effective RMS voltage and the tissue impedance is 1.65 MRayls (or 1.5 MRayls if water).

The two transmission functions, $Tx_ratio(f)$ and $TTF(f)$, derived here can be used to obtain spectral estimates for total acoustic power output, P_o , system efficiency η , transducer efficiency η_{xd} , radiation resistance R_a , and dissipation resistance R_d . The model prediction of acoustic output power for the entire array operating in phased parallel fashion is found as

$$P_o(f) = \frac{|TTF(f)|^2}{2 \cdot Z_{tissue}} \cdot V_{DUI}^2 \cdot DC \cdot N \quad (13)$$

where DC is the duty cycle defined as the burst length divided by pulse repetition interval, and N is the element count which is 64 for both the G3 and G4 designs, with a “single element” here considered to be two times that of an element of one of the two outside arrays, since they are operated in parallel.

Only the real parts of the output and input powers are considered here. It is reasonable to ignore the reactive power at the transducer since it is not of first order importance in the analysis, although in practice a high reactive power can add significantly to the net loading on the system power supply and should be considered due to power supply limitations. The efficiency characteristic of acoustic power delivered to the tissue for a given input power from the system is then defined to be the acoustic output power divided by the real part of the input power (Lamberti *et al.*, 1997), or

$$\eta(f) = \frac{\text{Re}\{P_o(f)\}}{\text{Re}\{P_i(f)\}} = \left[\frac{1}{2} \frac{|F_{tissue}|^2}{Z_{tissue}} \right] \cdot \left[\frac{1}{2} \frac{|V_{DUI}|^2}{|Z_{cab_{in}}(f) + R_s|} \cos(\angle(Z_{cab_{in}}(f) + R_s)) \right]^{-1} \quad (14)$$

which simplifies with the definition of the transmission transfer function to yield

$$\eta(f) = \frac{|TTF(f)|^2}{Z_{tissue}} \cdot \frac{|Z_{cab_{in}}(f) + R_s|}{\cos(\angle(Z_{cab_{in}}(f) + R_s))}, \quad (15)$$

where

$$Z_{cab_{in}}(f) = \frac{[MTCX(f)]_{0,0} \cdot Z_{tissue} + [MTCX(f)]_{0,1}}{[MTCX(f)]_{1,0} \cdot Z_{tissue} + [MTCX(f)]_{1,1}} \quad (16)$$

is the electrical impedance looking into the tuning inductor and cable at the system end.

Similarly the total acoustic power output and power efficiency of the transducer-only case can be defined, which is based on all the same assumptions, but without the system transmitter output resistance, the lossy tuning inductor, and lossy cable. If an ideal voltage transmitter with a zero source impedance drives the transducer directly, the following relationships result for this “transducer only” case. The transmit ratio and transmission transfer function for this case are given by

$$Tx_ratio_{xd}(f) = 1 / \left[MX(f) \cdot \begin{bmatrix} 1 & Z_{tissue} \\ 0 & 1 \end{bmatrix} \right]_{0,1} \quad (17)$$

$$TTF_{xd}(f) = \text{Force}_{tissue} / V_{DUI} = Tx_ratio_{xd}(f) \cdot Z_{tissue} \quad (18)$$

and from which the total acoustic power emitted from a transducer driven with an ideal voltage source can be expressed as

$$P_{o_xd}(f) = \frac{|TTF_{xd}(f)|^2}{2 \cdot Z_{tissue}} \cdot V_{ideal}^2 \cdot DC \cdot N. \quad (19)$$

The expression to show the efficiency in the conversion of real acoustic power out from real electrical power input from the ideal source is

$$\eta_{xd}(f) = \frac{|TTF_{xd}(f)|^2}{Z_{tissue}} \cdot \frac{|Z_{xd_{in}}(f)|}{\cos(\angle Z_{xd_{in}}(f))} \quad (20)$$

where

$$Z_{xd_{in}}(f) = \frac{[MX(f)]_{0,0} \cdot Z_{tissue} + [MX(f)]_{0,1}}{[MX(f)]_{1,0} \cdot Z_{tissue} + [MX(f)]_{1,1}}. \quad (21)$$

Since it is possible to have various peaks in the power output and efficiency spectra, it may of interest to find the product of the two to illuminate the best spectral band of operation. A normalization of this product can be arranged and expressed as the power optimum spectrum for the full “system,” or

$$P_{opt_s}(f) = [P_o(f) \cdot \eta(f)]_{norm} \quad (22)$$

as seen in shown in Figure 4 (1e and 2e) for both G3 and G4. A frequency dependent relationship of acoustic radiation resistance and loss (or dissipation) resistance for the transducer can also be derived. Since it is desired to maximize the thermal efficiency of the transducer design, the ratio of dissipation resistance to acoustic resistance should be maximized. If the acoustic power, P_a , delivered to the tissue load is considered the same as an equivalent electrical power which results from the transducer voltage across the electrical equivalent of the transducer acoustic resistance, R_a as shown in Figure 2, then

$$P_a(f) = (V_{xd}(f))^2 / R_a(f) = (\text{Force}_{tissue}(f))^2 / Z_{tissue} \quad (23)$$

and by rearranging terms and using the definition for the $TTF_{xd}(f)$

$$R_a(f) = 1 / \left[\left[(\text{Force}_{tissue}(f))^2 / Z_{tissue} \right] / (V_{xd}(f))^2 \right] = 1 / \left[|TTF_{xd}(f)|^2 / Z_{tissue} \right] \quad (24)$$

or simply

$$R_a(f) = \frac{Z_{tissue}}{|TTF_{xd}(f)|^2}. \quad (25)$$

This “parallel model electrical equivalent of the transducer acoustic resistance” is, for brevity, simply referred from this point forward as the transducer acoustic resistance, with our parallel model assumed. According to the definition of efficiency used in this work which is real power output for real power input, and in consideration of the parallel model arrangement of Figure 2, a relationship for frequency dependent dissipation resistance, $R_d(f)$, of the transducer itself can be formed. The voltage in the expression of real power out divided by real power input vanishes leaving the simple relation of transducer-only efficiency, $\eta_{xd}(f)$, equaling $R_d(f) / (R_d(f) + R_a(f))$. Rearranging, this becomes

$$R_d(f) = \frac{\eta_{xd}(f)}{1 - \eta_{xd}(f)} \cdot R_a(f). \quad (26)$$

In a manner similar to the rationale for the power optimal spectrum, the creation of another metric to reveal the best spectral band of operation in terms of the greatest power output with the smallest probe heating can be justified. The motivation here is to find the spectral peak of the transducer power available for the least heating independent of system power transmission issues. This relation (shown in Figures 4 (1f and 2f) and 5) may be formed by using the product of the power output spectrum of the transducer-only case and the ratio of the transducer dissipation resistance spectrum and transducer acoustic radiation resistance spectrum,

$$P_{\text{opt}_h}(f) = \left[P_{\text{oxd}}(f) \cdot \frac{R_d(f)}{R_a(f)} \right]. \quad (27)$$

2.4 Application of the Model with System Parameters for the G3 and G4 Designs

The KLM model discussed in the previous section was implemented using MathCAD (v11.0a, Mathsoft Engineering & Education, Inc.) and a standard desktop PC (Dell DIM4600, 2.8GHz Pentium, 1GB RAM). This modeling platform was used to gain the most flexibility for inclusion of all the loss parameters considered, and additionally because a multidimensional model was not essential due to expectations of reasonably low array cross-coupling and favorable array and aperture geometries. The model parameters shown in Tables 3 through 6 were used to obtain simulation results in Figures 4 and 5. Some minor adjustments in parameters such as k_{33} and layer velocities were made to better match model and laboratory results.

The results of a design comparison study between the G3 and G4 are shown in Figure 4 with every plot created with all 64x2 elements active, an effective 32.5% duty cycle, and a constant transmitter source voltage condition for each of the two probes. With 8.5V_{DUI} for the G3, and 14V_{DUI} for the G4 the spectra in 1a and 2a show equal power performance at f_0 (i.e. 1.5MHz) but with markedly different band shapes. The G4 design appears to have a higher total system efficiency at 58% and as well a transducer-only efficiency of 76%, both modeled at f_0 . The wide band nature of the G3 design depends greatly on the system transmitter and cable, since without these system components as shown in Figure 4 -1c the band shape collapses and the net power rises at f_0 to more than 3 times the full system case in 1a.

The comparison of Figure 4 (1f and 2f), and in Figure 5, shows that the G4 design provides greater acoustic power for given internal probe dissipation heat developed. This efficiency metric, P_{opt_h} , used here is the product of the acoustic output power (with the full system parameters considered) and the ratio of R_d and R_a as represented in (27). This is interpreted as the practical, whole system based efficiency for best power at minimal transducer heating. As shown in Figure 5, the chosen graphite composite thickness of 435 microns results in a peak operating frequency of 1.62 MHz which is very close to our imaging system operating frequency of 1.54 MHz.

3. Results

The two most recent multi-frequency probe designs, the G3 and G4, are fully completed probe assemblies and have been extensively examined in the laboratory. The only invasive measurements permitted, however, are the internal temperatures from thermistors placed in the backing material of the outside row arrays. Nonetheless, with the direct measurements of a) total acoustic power emitted, b) acoustic pressure at the beam focus and at the array face, and c) temporal impulse response wave shape, combined with the internal temperature measurements, a significant laboratory data set for direct comparison to model predictions can be compiled.

3.1 Modeling Verification

With care taken to operate the Antares imaging system below its maximum allowable power supply output at 1.54 MHz for the examination of the G3 and G4 low frequency outer arrays, the acoustic power transfer characteristics were measured with a radiation force balance (Ultrasound Power Meter, Model UPM-DT-1E, Ohmic Instruments Co., Easton, MD.) and are presented in Figure 6. The imaging system was operated in spectral Doppler Mode, with PRF at 5 kHz, and a duty cycle of 32.5%. The three piezoceramic layer transducer array (G3) and the single piezoceramic layer (G4) designs show acoustic power transfer characteristics of 28.9 and 10.9 in units of $\text{mW/V}_{\text{DUI}}^2$ respectively. The acoustic pressure transfer characteristic at the elevation focus region is 180 and 100 $\text{kPa/V}_{\text{DUI}}$ for the G3 and G4, respectively. As expected, the acoustic output power is a linear function of squared input voltage. Since the G3 is a triple layer design, the ideal ratio of the power transfer characteristic slopes of the two designs should be three squared (i.e. 3 layers with 3 times the power per layer) for the same drive voltage across each layer, but is significantly less which suggests the operation of the G3 is less efficient than the G4. This assertion is justified in accordance with the model predictions in Figure 4, comparing panels 1b and 2b.

The Antares imaging system offers direct selection and control of eleven transmission frequencies between 1.0 MHz and 2.5 MHz which can be of great use. Acoustic output power spectral measurements of the two low frequency outer arrays of fully assembled G3 and G4 probes at 1.0, 1.11, 1.25, 1.33, 1.43, 1.54, 1.67, 1.81, 2.00, 2.22, and 2.5 MHz, using the acoustic radiation force balance have been performed. The result of these discrete frequency measurements are shown in Figure 7 along with the KLM derived estimate of acoustic output power spectral response. For these laboratory measurements, the spectral Doppler mode DUI voltages used were 6 and 14 volts for the G3 and G4, respectively. The model prediction includes all the parameters in transmission including the system transmitter characteristics, the cable and the transducer stack itself with double and single matching layers for G3 and G4 respectively.

The G3 and G4 low frequency outer arrays were used in B-mode operation with the Antares imaging system to enable acquisition of the “impulse response” wave shapes measured in the near field (5 mm depth) of one array with a small hydrophone (Muller, Onda Corp., Sunnyvale, Ca.). To compare with the KLM model result, the frequency-transformed 1.5 cycle drive signal of the system transmitters was used as the model driving function. The results are presented in Figure 8.

The tissue heating performance of the CoLinear G4 design has been successfully assessed with direct temperature feedback in a murine model (Kruse *et al.*, 2007). These results are the subject of a companion work generated in the laboratory and are not discussed here for reasons of brevity.

3.2 Design Efficiency and the Real part of the Array Impedance

The efficiency metric used here is defined as the real acoustic power delivered to the tissue load divided by the total real power delivered to the array. Considering this definition there is no concern in this study with the reactive loading of the system transmission path and of the transducer itself; specifically, no optimization of the small series tuning inductor was made to help in the cancellation the capacitance of the transducer and transmission line, but this is certainly a concern for overall system optimization and best use of available power. Measurements can be made on completed probes to compare with modeled results, and most importantly, it is shown below how the acoustic and dissipative resistances can be inferentially measured and compared with model results.

3.2.1 Laboratory Estimation of Acoustic and Dissipation Resistance—Standard techniques were used for the measurement of acoustic power output and internal transducer heating on fully built and assembled G3 and G4 probes at the eleven available transmit frequencies. Laboratory based estimates were then generated of the two components of electrical motional resistance, R_a and R_d .

Both R_a and R_d can be estimated experimentally as relative values around a model prediction at the center frequency, f_0 , which is 1.54 MHz. For example, the R_a spectral resistance values can be assumed as linear scalars from the acoustic power measured in the laboratory. If the assumption that the voltage across the transducer is constant is reasonable, then the simple voltage and power relationship can be used to estimate the spectral values of R_a as a scalar of the measured output power at various frequencies, or

$$R_a(f) = V_{xd}^2 \cdot P_a^{-1}(f) \quad (28)$$

where V_{xd} is the voltage across the transducer and $P_a(f)$ measured acoustic output power. If V_{xd} is truly constant across the spectrum then the acoustic power measured alone can be used to scale the estimate of $R_a(f)$ from the model derived resistance at mid-spectrum $R_a(f_0)$. Enabled with a well verified model, this approach might be considered very reasonable. With V_{xd} constant across the spectrum, the estimates of the radiation resistance could be scaled as

$$[R_a(f)]_{\text{est}} = [R_a(f_0)]_{\text{model}} \cdot \left[\frac{P_a(f_0)}{P_a(f)} \right]_{\text{lab}} = \frac{V_{xd}^2(f)}{P_a(f)} = \frac{V_{xd}^2(f)}{P_{ac}(f)} \quad (29)$$

where the acoustic output power $P_a(f)$ is identically the power measured in the laboratory with the cable and transmitter in place. To identify this power as the “acoustic power with the system cable,” we use the variable name $P_{ac}(f)$.

3.2.2 Generation of a Transducer Load Correction Factor—Since the KLM model has shown excellent agreement in various laboratory measurements there may be justification for the use of $R_a(f_0)$ from model prediction as the scalable radiation resistance at f_0 . The G4 transducer design produces a transducer impedance very much higher than the sum of transmitter resistance and cable resistance and therefore reasonably obeys the assumption of a relatively constant V_{xd} at all frequencies of interest. The G3 transducer design however is quite different since the relatively low G3 transducer impedance near resonance creates a substantial loading effect on the cable and system transmitter. The transmission voltage across the G3 transducer is not a constant across the frequency spectrum, but this can be corrected by using a model derived scaling factor which can compensate appropriately for the varying V_{xd} across the spectrum.

The undesired transducer voltage variation across the spectrum can be eliminated where the ratio above is $V_{xd}^2(f)$ over the output power without the cable. Removing the voltage variability by re-scaling the first estimate for $R_a(f)$ by a normalized ratio of expected power spectra, the new estimate becomes

$$R_a(f) = [R_a(f)]_{\text{est}} \cdot \frac{P_{ac}(f)}{P_{aNC}(f)} \cdot \frac{P_{aNC}(f_0)}{P_{ac}(f_0)} \quad (30)$$

where $P_{aNC}(f)$ is the calculated acoustic power without the system loading effect (i.e. where the source voltage is identical to the transducer voltage). So inserting (29) into (30) one may obtain a reasonable estimate of resistance as

$$R_a(f) = [R_a(f_0)]_{\text{model}} \cdot \left[\frac{P_{aC}(f_0)}{P_{aC}(f)} \right]_{\text{lab}} \cdot \left[\frac{P_{aC}(f)}{P_{aNC}(f)} \cdot \frac{P_{aNC}(f_0)}{P_{aC}(f_0)} \right]_{\text{model}} \quad (31)$$

where the spectral powers of $P_{aC}(f)$ and $P_{aNC}(f)$ are the previously model derived variables of $P_o(f)$ (13) and $P_{o_xd}(f)$ (19) respectively. The model-derived third term in (31) is the normalized squared voltage correction factor and is plotted in Figure 9 to demonstrate the contrast in transducer loading between G4 and G3. It is evident that the high impedance G4 design produces a relatively flat spectrum close to unity as expected and the G3 is highly varying across the band of interest.

3.2.3 Laboratory and KLM Model Comparisons of Resistance Estimates—

Laboratory data were collected and are shown in Table 7 in the P_a and ΔT columns for both G3 and G4 probes; all the resistance values are calculated. These data were collected with both the G3 and G4 completely assembled probes mounted in the acoustic radiation force balance with the Antares imaging system adjusted to operate in spectral Doppler mode at a PRF of 5000 and a duty cycle equal to 32.5%. To maintain the same duty cycle throughout the tests at the 11 different frequencies available, the number of transmit cycles were adjusted for the frequency in the test. The comparison of model derived and laboratory sample data resulted in Figures 10 and 11.

Since it is simply not possible to directly measure the spectral R_a and R_d values for a completely assembled probe, the use of lab data in combination with model “seed” values at f_0 (shown in bold in Table 7) and model derived loading correction factors previously discussed enable an inferential assessment of these important resistances and permit comparison to the KLM model results.

For the calculation of R_a for both G3 and G4, the power measured at a given frequency was used to scale the model derived “seed” estimate found at f_0 with (25). This intermediate value shown in Table 7 as R_a est. was then corrected to obtain the best estimate of the actual transducer voltage as in (31) and the result is shown in the table as the scaled laboratory estimate of R_a . These discrete lab based estimates are plotted against the KLM model derived curve for R_a shown in Figures 10 and 11.

The temperature data in Table 7 were collected by using a small thermistor (BetaTherm, 0.5mm diameter, p/n 2K7MCD1) placed during assembly in the backing material within 2 mm of the piezoceramic of one of the low frequency outer row arrays. The temperature elevation measurement for the two probes at each frequency was measured at 30 seconds following the initiation of acoustic power production.

The calculation of R_d is performed by a method similar to the description of R_a discussed above. The rationale is based on the direct linear relationship between a short duration temperature increase due directly to the heat of power dissipation. This dissipation power is inversely proportional to the effective electrical dissipation resistance, R_d . Thus, a first estimate of R_d can be made based on the inverse scaling of the model estimated resistance at f_0 . Since the same spectral voltage variation will occur for our R_d estimation, the scaling factor in (31) can be applied and the result is shown in Table 7 as the scaled laboratory estimate of R_d . These discrete lab based estimates are plotted against the KLM model derived curve for R_d shown in Figures 10 and 11.

3.2.4 Dissipation Resistance Estimates and Internal Probe Heating—Internal

temperature measurements, made with the thermistors placed in the low frequency array backing material, were made for both G3 and G4 at various acoustic power output levels to determine their performance during simulated use in tissue heating experiments. As shown in

Figure 12, the G4 internal temperature rise in continuous operation at 4 Watts of acoustic output is only 10.5 °C at 100 seconds which is superior to the G3 performance operating at half that acoustic output. The result here follows the relationship of R_d/R_a for each probe; the higher the ratio the better the general acoustic output against internal rise in temperature.

4. Discussion

A clear relationship has been demonstrated between the KLM model derived acoustic and dissipation resistances, R_a and R_d , and the acoustic power output and thermal response, respectively, for both the G3 and G4 low frequency array designs. The ability to derive reasonable estimates of these resistances is important for not only high efficiency designs, but can be important also in the optimization of thermal noise resistance, which should be strongly linked to the R_d estimate. In many designs, the fundamental limit in device sensitivity is largely dependent on this thermal noise resistance of the piezoceramic transducer.

The estimate of R_a , derived and verified in the laboratory study, appears to be reasonable considering the direct measurement of acoustic output power and high confidence in the actual voltage across the transducer. To be complete here, an effort should be made to account for all the power loss within the probe head and thereby enable the direct confirmation of our derived estimate for the dissipation resistance, R_d , as the KLM predicts. This important study is the subject of work currently underway.

As shown in Table 8 and Fig. 4-1b, the G3 design requires a large system power delivery capability as compared to the G4 for the same acoustic power developed. Further, the reduction in array dissipation with the G4 design results in longer available working time to heat tissue in the 100 seconds required. The G3 inefficiency in power use results in much of this unused power spent in heat production in the system transmitters and cable resistance. The G3 design however has been useful for applications where short bursts of high pressure are needed, e.g. in the case of microbubble destruction or microvessel permeability enhancement. The G4 is designed, by contrast, to operate in a much longer burst mode which is very important for lower pressure mild hyperthermia applications.

4.1 Transducer Resistance by Analytic Means

To lend confidence in our methods to determine good estimates of R_a and R_d , it is of great interest to compare our model derived estimate for the parallel model equivalent electrical resistance against an adequate analytic formulation. Several common analytical approaches are examined here in contrast to the KLM derivation of the electrical model parallel sum.

4.1.1 Motional Resistance by KLM Model Method—The relationship can be established between the net motional electrical resistance for our parallel model with the previously derived KLM values for $R_a(f)$ and $R_d(f)$ by simply taking these two resistances in parallel, which is the “parallel, KLM” resistance, or

$$R_{pk}(f) = \frac{R_a(f) \cdot R_d(f)}{R_a(f) + R_d(f)} \quad (32)$$

and is plotted for G3 and G4 in Figures 14 and 15, respectively.

4.1.2 Motional Resistance by Conventional Derivation of Transducer Electrical Impedance—To proceed with a confirmation of results, an effort can be made to find a “best” analytical estimation approach, as well as other approximations with which to compare against the KLM and laboratory results.

It can be assumed that the acoustic impedance for the backing layer, Z_B , and for the ceramic itself, Z_C , in the G3 and G4 designs can be simply described as static magnitudes to permit some reasonable ease in calculations to be described here. Using the standard transmission line formula (33) N times for N matching layers, the frequency dependent acoustic impedance can be expressed as it appears at the front surface of the piezoceramic core as

$$Z_1(f) = \frac{Z_2(f) \cos(k_{m1} t_{m1}) + j Z_{1s} \sin(k_{m1} t_{m1})}{Z_{1s} \cos(k_{m1} t_{m1}) + j Z_2(f) \sin(k_{m1} t_{m1})} \cdot Z_{1s} \quad (33)$$

which includes the material losses and frequency dependencies in each of the matching layers. Figure 13 shows the matching layer ordering considered here. In the computation of $Z_1(f)$ it can be seen that $Z_2(f)$ is the complex impedance of the last layer computed, Z_{1s} is the static magnitude of the first (inner) matching layer, k_{m1} is the wavenumber $2\pi/\lambda$ for the first matching layer, and t_{m1} is the thickness of that layer. Both the G3 and G4 designs use a soft, flat polyurethane lens with a velocity characteristic close to water; this justifiably benign layer, used only for mechanical protection of the arrays, is omitted from examination as an acoustically active matching layer.

Thus, for a given design and our estimates for $Z_1(f)$, Z_B , Z_C , transducer array electromechanical coupling k_{33} , and transducer bulk capacitance, a computation can be made to approximate the motional resistance R_m for the Van Dyke tank circuit model form where the bulk capacitance is in parallel with the series R_m , L , C elements. This R_m is considered here to be the “parallel” tank model resistance for a piezoelectric transducer.

As developed by many before (Desilets *et al.*, 1978; Kino, 1987; Goll, 1979), the motional electrical equivalent resistance can be found by first using the well known expression for the electrical impedance looking into the two thin electrodes that form the piezoelectric ceramic front and back plate surfaces. This electrical port is described by a 3 X 3 impedance matrix relating forces, velocities, voltage, current and transmission parameters at the transducer layer. The analytic impedance is of interest, or the ratio of voltage to current at this layer which is expressed by

$$Z(f) = \frac{1}{j2\pi f C_0} \left[1 + \frac{k_{33}^2}{\beta_{xd} t_{xd}} \frac{j Z_C (Z_1(f) + Z_B) \sin(\beta_{xd} t_{xd}) - 2 Z_C^2 (1 - \cos(\beta_{xd} t_{xd}))}{(Z_C^2 + Z_1(f) Z_B) \sin(\beta_{xd} t_{xd}) - j Z_C (Z_1(f) + Z_B) \cos(\beta_{xd} t_{xd})} \right] \quad (34)$$

where C_0 is the bulk transducer capacitance, k_{33} is the array element electromechanical coupling, β_{xd} is the PZT propagation constant as a function of frequency, t_{xd} is the transducer thickness, and $Z_1(f)$ is the complex and frequency dependent acoustic impedance load on the front side of the ceramic, and Z_B , Z_C , as mentioned earlier are treated as single value magnitudes.

It is not easy to write an accurate, single expression in compact form that represents the motional resistance desired, since to attempt this with modest accuracy in mind, the calculation needs to retain firstly the sense that $Z_1(f)$ is a complex quantity which makes the process somewhat difficult. Moreover, to proceed, one must use apply (33) N times for N matching layers, followed by the application of (34) to find $Z(f)$, and finally by taking the appropriate real part of $Z(f)$. For the “parallel” model, this last step actually requires the inverse of the real part of the admittance, or equivalently

$$R_{pa}(f) = \left[\text{Re} \left[Z(f)^{-1} \right] \right]^{-1} \quad (35)$$

and is plotted in Figures 15 and 16, for G3 and G4 respectively.

$R_{pa}(f)$ is reasonably precise and depends largely on the accuracy of the input parameters used to carry out the calculation, especially the knowledge of the $Z_1(f)$ acoustic impedance which is frequency dependent, complex and must be incorporated into the computation prior to the segregation of the real from the imaginary part in the computation.

4.1.3 Motional Resistance Estimate with Acoustic Loading Static at f_0 —For comparison purposes, a less accurate but perhaps more easily calculated result for the R_p may be attempted. In this case the acoustic impedance load $Z_1(f)$ may be considered to be simply one complex value at the center frequency, or $Z_1(f_0)$. This evaluation is considered to be

$$R_{pas}(f) = R_{pa}(f) \Big|_{Z_1(f)=Z_1(f_0)} \quad (36)$$

and is plotted in Figures 14 and 15, for G3 and G4 respectively.

4.1.4 Motional Resistance Estimate with a Single Expression—A further simplification for comparison may be established for the computation for the “parallel model equivalent electrical resistance” with the well known assumption which uses the conditions of a) $\beta t_{xd} = \pi$, and b) $f = f_0$. These generally are contradictory requirements, but if the former is chosen as the primary requirement which dictates the frequency chosen (where the ceramic thickness is $\lambda/2$ in the fundamental resonance region, $f_{\lambda/2}$), it can be shown that the trigonometric terms of (34) simplify. If the use of static magnitudes for Z_1 , Z_B , and Z_c can be assumed, and as well the use the frequency where the wavelength in the ceramic is $\lambda/2$ (although $f_{\lambda/2} \neq f_0$), it can be shown that the approximation becomes

$$R'_{pas} = \frac{1}{2\pi f_{\lambda/2} C_0} \cdot \left[\frac{\pi (Z_1 + Z_B)}{4k_{33}^2 Z_c} + \frac{4k_{33}^2 Z_c}{\pi (Z_1 + Z_B)} \right] \quad (37)$$

and is plotted as a single (square box) point in Figures 14 and 15, for G3 and G4 respectively.

This result is in conflict with the generally accepted formulation for the parallel model resistance (Kino, 1987; Whitworth, 2001; Cobbold, 2007; Szabo, 2004); the difference is the second term which is absent in the standard published form, but is certainly not negligible for currently used array designs with coupling coefficients greater than 0.5. For a typical ratio of $Z_c/(Z_1 + Z_B)$ of 1 to 3, the second term does become small (10% or less) for coupling coefficients less than 0.3.

To contrast this method with our earlier assessments of motional resistance, the single term version of R'_{pas} is shown in Figures 14 and 15, plotted as an “X.” These estimates assume scalar values for k_{33} , Z_1 , Z_B , and Z_c as: 0.64, 11.2 MRayls, 3.2 MRayls, and 30.8 MRayls, and 0.60, 21 MRayls, 3.2 MRayls, and 30.8 MRayls, for the G3 and G4 designs respectively.

4.2 Transducer Resistance Comparison between the G3 and G4 Designs

For both the G3 and G4 designs there is good agreement between the KLM derived parallel model motional resistance, R_{pk} , and the well known analytic expression which is expressed as R_{pa} . This latter expression benefits from the acoustic loading impedance (i.e. $Z_1(f)$) calculated as a complex, frequency dependent variable at every frequency in the band of interest. The R_{pas} (i.e. static Z_1) estimate though assumes only a magnitude value of Z_1 at f_0 , and therefore shows good agreement at f_0 , but with only modest agreement in spectral shape. The simplified expression for the parallel model motional resistance is just an estimate at best. It is curious to note that the (incorrect) single term form for this expression is actually closer to the best estimators, R_{pk} and R_{pa} , at the piezoceramic half wavelength frequency. Nonetheless, complex

spectral response designs such as the G4, which are not uncommon for sonotherapeutic transducers, really require a more sophisticated approach for accurate computation.

The spectral estimation plots of parallel model motional resistance can be summarized as shown in Table 9. The KLM modeling result of (32) and the analytic impedance of (34) are considered to be the best estimators, and they agree generally across our spectrum of interest and show reasonably small differences at the center frequency, f_0 .

5. Conclusions

In comparison with its precursor design, the graphite composite used in the front matching layer design of the G4 probe has successfully decreased probe heating and increased power efficiency to permit the G4 to be used in laboratory investigations for mild hyperthermia applications. For application periods of 100 seconds, the G4 can produce 4 Watts of total acoustic power continuously with only a modest rise in internal probe temperature.

The study of thermally efficient probe design has also lead to the following important conclusions:

- a. transducers for confocal ultrasound imaging and therapy can be designed for a transducer optimized for thermal therapy differing from mechanical therapy,
- b. an unmodified conventional imaging system can provide the necessary power, together with a well designed probe, for mild hyperthermia applications,
- c. for thermal therapy, a graphite composite layer can improve the design of an acoustic transducer by lowering significantly the thermal resistance in the heat loss path, and with only modest reductions in bandwidth,
- d. the KLM model can offer methods for the direct computation of the electrical equivalent acoustic resistance, R_a , and the dissipation resistance R_d ,
- e. the well known analytic expression for transducer electrical impedance is only suitable to predict the combination motional resistance (which is the combination of R_a and R_d), and is only accurate across the band of interest if one uses the complex, frequency dependent form of the acoustic impedance which appears at the front port face of the internal ceramic resonator, and
- f. the simplified and commonly used single expression estimate of R_m is not very accurate in the prediction of the true motional resistance at the transducer center frequency.

Acknowledgements

Ed Gardner, Lisa Fearn, Bill Beck, Tom Clary, Levin Nock, and Pat Sutcliffe, in working with Siemens Medical Solutions and Siemens Corporate Research, are gratefully acknowledged for their valuable support in the development of the G2, G3 and G4 CoLinear probes and in DUI control parameters. Shengping Qin and Charles Caskey of the Department of Biomedical Engineering at the University of California, Davis are also acknowledged for their considerable contributions in reviewing the manuscript and in laboratory investigations. NIH CA 112356 and 103828 supported the development of the multi-frequency arrays.

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Figure 1a.

The multi-array CoLinear common features are two 6mm elevation outside arrays at 1.5MHz and a single 4mm elevation 5.3MHz array (a). The G3 and G4 are different in their outside row designs. The G3 uses a triple layer piezoceramic and two matching layers (b), while the G4 design uses a single layer piezoceramic and single matching layer (c).

Figure 1b.

A representative B-mode image acquired during a sonotherapy treatment of a mouse tumor model as a demonstration of the high level of image quality from the use of the 128 element center imaging array in each of the CoLinear probes. In this particular set up, the Antares system, operating as well in spectral Doppler with the low frequency array pair to produce the mild hyperthermia, is in closed loop control with thermocouple feedback with Doppler-PRF modulation to directly maintain a desired target temperature in the range of 41 to 43 °C.

Figure 2.

A generalized system interface for a transducer array. The top panel shows the two components of the real power delivered to the array elements as the desired acoustic (P_a) and undesired dissipative (P_d) power. The lower panel shows the principal loss components in power transmission with a simplified tank circuit model at resonance.

Figure 3.

The matrix elements of the KLM model used to characterize the CoLinear designs. A general model for a 2-way entire network is shown in (a), our CoLinear design for 1-way system transmission and tissue loading characteristics in (b), and details of the constituents are shown in (c) as a series of 2X2 matrices.

Figure 4.

The KLM model calculated spectral response comparisons for the G3 (1a – 1f) and G4 (2a – 2f) low frequency transducer designs, with computational results at f_0 for each case. All paired cases here are performed with unvaried transmit drive voltages (8.5V and 14V, for G3, and G4 respectively) needed to produce 2.1 Watts total acoustic power output with full system and cable. The 1a-2a power spectra with full system and cable pair are derived from (13); the power spectrum pair, 1c-2c, with an ideal voltage source driving the transducers are derived from (19). The efficiency spectra shown in 1b-2b and 1d-2d are derived from (15) and (20) respectively. The “power spectrum efficiency metric” plots of 1e-2e and 1f-2f are derived from (22) and (27) respectively.

Figure 5.

The transducer design efficiency (P_{opt_h}) summary spectra for five thicknesses of the graphite composite matching layer in the G4 design, and a single plot comparison for the G3 design with its double layer matching design.

Figure 6.
The linear characterization of the two CoLinear designs showing total actual laboratory acoustic power transmitted for a given driving voltage squared.

Figure 7.

The acoustic power output spectra, for the G3 design (top) and the G4 design (bottom). The KLM model prediction for power output is stated for $f_0=1.54\text{MHz}$, and as a solid line (see also Fig. 4-1a and Fig. 4-2a); the lab results are shown as points with 10% error bars, and with G3 operating as at a conservatively low power level in this test.

Figure 8.
The wave shape comparisons of model (dashed) and lab measurements (solid) of B-mode transmissions for the G3 design (top) and the G4 (bottom).

Figure 9.

The normalized, unitless correction factor values at the eleven test frequencies used for G3 and G4 estimates of R_a and R_d resistances. The correction for G4 is near unity due to very light loading on the cable and imaging system transmitters.

Figure 10.

The acoustic radiation resistance (R_a) and the dissipation resistance (R_d) for the G3 probe design. The solid and dashed traces are the model predictions for R_a and R_d respectively. The circles and crosses are calculated from lab data based on acoustic power measurements and thermistor step responses in long burst transmission tests. The boxed points at 1.54MHz are unique “seed” points that served as a model-derived reference points for absolute scaling of the lab data.

Figure 11.

The acoustic radiation electrical resistance (R_a) and the dissipation electrical resistance (R_d) for the G4 probe design. The solid and dashed traces are the model predictions for R_a and R_d respectively. The circles and crosses are calculated from lab data based on acoustic power measurements and thermistor step responses in long burst transmission tests. The boxed points at 1.54MHz are unique pivot points that served as model-derived reference points for absolute scaling of the lab data.

Figure 12.

The backing material embedded thermistor temperature response curves for the G3 and G4 designs at various acoustic outputs. Starting at 23 °C the G4 shows only a 10.5 °C rise at 100 seconds of 4W continuous output.

Figure 13.

The acoustic impedances Z_1 and Z_2 which are the loads for their corresponding interfaces for the direction shown.

Figure 14.

Comparison for the analytical estimates of equivalent motional resistance for the G3 design model. The KLM model was used to estimate R_{pk} , the parallel combination of R_a and R_d . The standard analytical estimate for motional resistance, R_{pa} , is presented along with the modified version, R_{pas} , and two simplified single expression estimates.

Figure 15.

Comparison for the analytical estimates of equivalent motional resistance for the G4 design model. The KLM model was used to estimate R_{pk} , the parallel combination of R_a and R_d . The standard analytical estimate for motional resistance, R_{pa} , is presented along with the modified version, R_{pas} , and two simplified single expression estimates.

Table 1

Array Parameters

Array Parameter	Inner Array	Outer Array(s)
Array Frequency	5.24 MHz	1.5 MHz
Array Element Count	128	64 elements in each array, operated in parallel
Array Apertures	4mm x 38.4mm	6mm x 38.4mm for each of two arrays
Array Element Pitch	1.01 λ	0.61 λ
G3: PZT layers	1	3
G3: Matching Layers	2	2
G4: PZT layers	1	1
G4: Matching Layers	2	1

Table 2

Notation and selected nominal values used in the KLM model

Symbol	Definition	Values/Units
A_{xd}	area of an outside row array element	m^2
C_{cab}	cable capacitance per meter	Farads m^{-1}
C_0	bulk transducer capacitance	Farads
DC	duty cycle or product of burst length and PRF	0.325
f_0	outer array frequency at bandwidth center	1.54MHz
f_m	frequency for a particular acousto-mechanical layer	sec^{-1}
f_t	test frequency for inductor	sec^{-1}
$Force_{tissue}$	acoustic force exerted on tissue medium	Newtons
k_{33}	array electro-mechanical coupling coefficient	
L	tuning inductor nominal inductance	Henrys
L_{cab}	cable series inductance per meter	Henrys m^{-1}
N	number of elements	64
P_a	acoustic power output from array	Watts
P_d	dissipative power from array	Watts
PRF	pulse repetition frequency	5000 Hz
Q_L	quality factor rating of inductor at f_t	
Q_m	material quality factor for a particular acousto-mechanical layer	
R_a	transducer (parallel model) acoustic radiation resistance	Ohms
R_d	transducer (parallel model) acoustic dissipation resistance	Ohms
R_m	transducer (parallel model) motional resistance	Ohms
R_s	system transmitter series source resistance	Ohms
R_{cab}	cable series resistance per meter	Ohms m^{-1}
R_L	electrical tuning inductor resistance	Ohms
$\tan\delta$	loss tangent for the piezoceramic	0.017
t_{xd}	piezoceramic thickness	m
V_{DUI}	system transmitter voltage selected by the operator of the diagnostic user interface (approximately V_{RMS} due to square wave nature of the Antares transmitter output)	Volts
vel_m	material velocity for a particular acousto-mechanical layer	$m sec^{-1}$
V_{ideal}	a transmitter voltage from a source without an internal series impedance, i.e. ideal	Volts
V_{xd}	transducer voltage	Volts
Z_B	acoustic impedance of the backing material	MRayls
Z_C	acoustic impedance of the piezoceramic	MRayls
Z_{cab}	electrical cable impedance	Ohms
Z_1	acoustic impedance load on the front side of the MRayls piezoceramic, looking into the first matching layer (may be either a function of frequency or as a static magnitude at $f_{\lambda/2}$)	MRayls
Z_2	acoustic impedance load on the front side of first matching layer towards the second layer	MRayls
Z_{KLM}	acoustic impedance used in KLM model	MRayls m^2
Z_{layer}	acoustic impedance for a particular layer	MRayls
Z_L	electrical tuning inductor impedance	Ohms
Z_{tissue}	acoustic impedance of animal tissue medium	1.65 MRayls
α_{layer}	frequency dependent real part of the acoustic propagation constant for a particular layer	$Np m^{-1}$
β_{layer}	frequency dependent imaginary part of the acoustic propagation constant for a particular layer	m^{-1}
β_{xd}	imaginary part of the piezoceramic propagation constant	m^{-1}
γ_{cab}	electrical cable propagation constant	m^{-1}
γ_{layer}	$\alpha_{layer} + j\beta_{layer}$, the complex acoustic propagation constant for a particular acoustic layer	m^{-1}
ϵ	dielectric constant of piezoceramic material	$F m^{-1}$
ϵ_r	relative dielectric constant of piezoceramic	
ϵ_0	dielectric constant of free space	$8.85 \cdot 10^{-12} F m^{-1}$

Table 3

Transducer Piezoceramic

Parameter	Unit	G3	G4	Comments
Elevation length	mm	6 (x2)	6 (x2)	two 6mm elevation elements
Width	mm	0.56	0.56	0.6 mm array pitch
Thickness	mm	1.02	1.02	approximately same for both
Velocity	m/sec	3480	3780	triple will be 10% slower
Relative Dielectric		20,880	2320	triple is 9 times single layer
Mechanical Q		25	50	triple layer is about half (Zipparo, 2003)
k_{33}		0.60	0.64	triple is lower due to flex
Acoustic Impedance	MRayls	30.8	30.8	
$\tan \delta$		0.017	0.017	

Table 4

Transducer Backing

Parameter	Unit	G3	G4	Comments
Thickness	mm	10	10	estimate
Acoustic Impedance	MRayls	3.2	3.2	
Velocity	m/sec	1850	1850	
Attenuation	dB/mm	5.6	5.6	estimated, at 1.54MHz

Table 5

Transducer Matching Layers

Parameter	Unit	G3	G4	Comments
Thickness	mm	0.53	0.435	1 st matching layer thickness and impedance agree with DOE optimization method (McKeighen, 1998)
Acoustic Impedance	MRayls	8.11	5.9	
Velocity	m/sec	2500	2630*	* function of freq, neg. slope at 1.54MHz
Attenuation	dB/mm	0.58	0.32	
Thickness	mm	0.36	---	2 nd matching layer thickness and impedance agree with DOE optimization method (McKeighen, 1998)
Acoustic Impedance	MRayls	2.57	---	
Velocity	m/sec	2381	---	estimated, at 1.54MHz
Attenuation	dB/mm	1.13	---	

Table 6

System and Cable

Parameter	Unit	Description	Comments
Transmitter type		bipolar	
Transmitter Rout	ohms	32	estimated, load dependent
Tuning Inductor	μH	4.7, 3.3	series inductor for G3, G4; QL = 40 @ 7.9MHz
Cable type		90 ohm coax	Z0 nominal is 90 ohms
Cable length	m	2	approximate length
Cable resistance	Ω/m	9.9	44AWG (50.8 μm in dia.)
Cable inductance	$\mu\text{H}/\text{m}$	0.47	
Cable capacitance	pF/m	59	

Table 7
Power and Thermistor Data for Array Parameter Estimation

Freq.	Cycles	G3					G4				
		P _a (W)	ΔT (°C)	R _d est.	R _d (Ω)	R _a est.	R _a (Ω)	P _a (W)	ΔT (°C)	R _d (Ω)	R _a (Ω)
1	65	1.01	1.82	211	421	150	300	0.64	2.86	3136	4020
1.11	72	0.99	2.02	190	374	154	302	1.80	7.35	1219	1420
1.25	81	1.04	2.19	175	298	146	248	2.84	5.76	1555	900
1.33	86	1.04	1.99	193	300	146	227	2.42	3.97	2256	1058
1.43	93	1.03	2.51	153	190	148	184	1.95	2.41	3712	1312
1.54(f ₀)	100	1.04	2.30	167	167	146	146	2.10	2.30	3896	1218
1.67	108	0.97	2.17	177	199	157	175	2.71	3.80	2361	944
1.81	118	0.86	2.47	156	301	177	342	3.23	6.82	1315	791
2	130	0.64	3.51	110	453	237	980	1.10	2.34	3833	2317
2.22	144	0.15	1.59	242	1075	1008	4476	0.25	1.59	5651	10317
2.5	162	0.05	0.49	788	1806	3295	7548	0.10	0.97	9200	25415

Table 8
Design Efficiency Comparison between G3 and G4 at f_0

Probe Design Type	LF Array Design: PZT layers, match. layers (Layer Count)	Calculated Total System Power-Sourced (Watts)	Array Acoustic Power (Watts)	Calculated Array Dissipation Power (Watts)	Time for 15 °C rise of internal temperature at $P_a = 2.5W$ (sec.)
G3	3, 2	11.4	2.5	2.18	~ 200
G4	1, 1	4.3	2.5	0.78	> 1200

Table 9
Motional Resistance Estimate Comparisons from Figs. 14 and 15

Modeling Method	Parallel Model Motional Resistance Estimates				
	G3		G4		
	$R(f_0)$ (Ω)	error %	$R(f_{1/2})$ (Ω)	$R(f_0)$ (Ω)	error % $R(f_{1/2})$ (Ω)
KLM	80		85	928	547
Analytic Impedance	67		92	968	634
KLM & Analytic Difference at f_0		16.5			- 4.1
2-term Simplified Expression			194		1294
1-term Simplified Expression			70		523