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Method for transducer transient suppression. I: Theory

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Method for transducer transient suppression. I: Theory

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The problem of driving a transducer in such a way as to produce a tone burst of steady-state sound radiation in the surrounding fluid medium is considered. The goal is to determine the driving voltage waveform to apply to a transducer to produce an acoustic pressure waveform in the fluid that is a segment of a steady-state sine wave, beginning and ending at zero crossings of the sine, i.e., the usual turnon and turnoff transients are suppressed. The theoretical driving voltage waveform for a spherical transducer is shown to consist of a sum of a pedestal voltage, a ramp voltage, and a sinusoidal voltage that is phase shifted with respect to the sinusoid appearing in the fluid. Both theoretical and numerical calculations are given here. The following paper presents results of experimental measurements. The measurements were carried out on several spherical transducers (one of which was selected for presentation) and on an array of piezoelectric tubes. These experiments confirm the validity of the theory.

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INTRODUCTION

The problem of driving a transducer in such a way as to produce no transients in the sound radiated into the surrounding fluid medium is considered. The solution of this problem is of interest for two reasons. First, in order to calibrate a transducer under environmental conditions of temperature and pressure, such calibration must often be effected in a test facility of rather limited size. If the transient portion of the pressure waveform produced immediately after the driving voltage is applied to the transducer is of sufficiently great duration, as is often the case, reverberations from the boundaries of the test facility can interfere with the direct radiation from the transducer and can preclude accurate calibration.

The second area of interest is in scattering or reflection measurements that might also be carried out in a test facility of limited size. In this case, in order to separate the interrogating radiation from the scattered or reflected radiation, it is necessary to place the sound detector at a sufficiently great distance from the scatterer that the turnoff transient has diminished the interrogating radiation to a level that is well below that of the scattered radiation at the time of its reception. If the rate of decay of the interrogating radiation after drive turnoff is sufficiently slow, the required detector-to-sample separation may have to become great enough that the facility boundaries again interfere with the desired measurement.

The method used here to solve the transient problem, which we will refer to as the "transducer transient suppression" method, involves analytically evaluating an equivalent circuit for the transducer of interest. Thus, the transient-suppressing driving voltage waveform deduced here will actually be that which is appropriate for driving the *circuit* in a transient-suppressed mode. The success of the method in suppressing transient radiation from the transducer will hinge upon the fidelity of the equivalent circuit to the transducer of interest.

The problem of interest could alternately be approached from the point of view of transfer function theory.[1-4] (Other methods of pulse shaping are based on feedback[5] or the use of a Wiener filter.[6]) That is, if the transfer function of a transducer can be accurately determined, appropriate Fourier Transforms can (in principle) be evaluated to determine how to drive the transducer in any desired way. However, since this method involves the use of deconvolution, it is highly sensitive to noise. The equivalent circuit approach is not expected to be particularly noise sensitive. The equivalent circuit approach also has the advantage that the parameter values can be restricted by *a priori* knowledge of the behavior of electric circuit elements. For example, the circuit resistances obviously must be positive, and the capacitances and inductances must be real valued. In the transfer function approach, a completely unphysical model might be numerically generated due to the influence of noise, or due to other effects that might be unknown to the experimenter. It is demonstrated that the equivalent circuit approach also helps in selecting a power amplifier that is best suited to achieving transient suppression.

In Sec. I, the transient suppression method is developed using the simple LCR circuit. While much of what is presented here is well known, this case is helpful to consider since it gives insight into the calculational method. The analysis of this case is also useful for gaining an understanding of why the method works in the more complicated cases of practical interest. The method also happens to produce an exact solution for the LCR circuit. The application of the method to a spherically shaped transducer is considered in Sec. II. Included is a discussion of the approximate nature of the solution, as well as a discussion of what is required to obtain the exact solution of the relevant differential equation. This exact circuit solution is useful in evaluating the accuracy of the approximate transient-suppressing drive that is presented. In Sec. III numerical experiments that were carried out to investigate the effectiveness of the ap-

proximate theoretical solutions are described. Section IV presents the transient-suppressing drive in a convenient mathematical form, and also gives a discussion of the drive's various features. Section V gives a summary, the conclusions, and a description of future work.

I. LCR CIRCUIT

A. Background

To illustrate the transient-suppression concept, we will consider first the simple LCR circuit, driven by an arbitrary-waveform (ARB) voltage source which produces the arbitrary waveform $V(t)$ (see Fig. 1). We consider this circuit not because it represents a realistic equivalent network for any transducer of interest, but rather because it is the simplest case that contains all the salient features of a damped oscillating system.

Any excitation of an oscillating system tends to cause the system to oscillate at its resonance frequency (or frequencies). For example, in the case of the LCR circuit of Fig. 1, if the ARB were replaced by a "dead short" (i.e., a section of conducting wire), and if the capacitor had some initial nonzero charge then, when switch S is closed, the voltage waveform observed across the resistor would be a damped sinusoid. This damped sinusoid is characterized by the circuit's resonance frequency

$$f_0 = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}$$

and an exponential damping factor $e^{-Rt/2L}$, where t is the time since the switch was closed.

Next, suppose the capacitor C is initially uncharged and that the ARB is driven to suddenly apply a gated sinusoidal voltage to the circuit at the moment switch S is closed. (The frequency f_0 of this driving waveform need not be equal to the resonance frequency of the circuit.) In this case, the voltage waveform observed across R would appear to be the sum of two waveforms. One of these waveforms would simply be a steady-state sinusoid characterized by the frequency f_0 of the driving waveform. The other waveform would be the same type of damped sinusoid as described above. After the termination of the driving voltage sinusoid, the voltage waveform observed across R would only be the damped sinusoid associated with the circuit's resonance frequency.

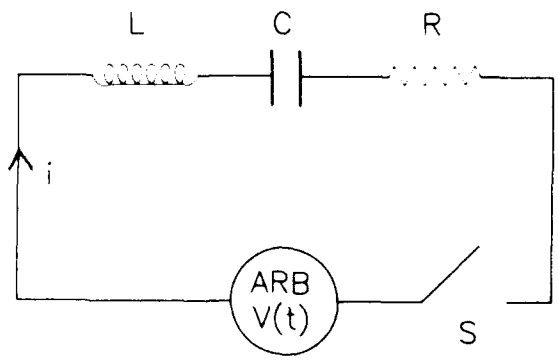

FIG. 1. An LCR circuit. The symbol ARB denotes an arbitrary waveform generator that produces the voltage waveform $V(t)$.

The transient observed near the start of the voltage waveform appearing across the resistor is termed a "turnon transient." The transient observed across the resistor after the cessation of the drive is a "turnoff transient" or "ring-down transient." We seek here a modified driving voltage waveform so that both of these transients are suppressed. Of course, there is no *a priori* reason to expect that a drive can be found that satisfies the desired 100% transient suppression requirement. However, in the case of the LCR circuit such a drive *can* be found. In the case of a more realistic transducer equivalent network, as considered in Sec. II, transient suppression is only approximately achieveable.

B. The solution

The differential equation of the LCR circuit of Fig. 1 is

$$L\frac{di}{dt} + iR + \frac{q}{C} = V(t), \tag{1}$$

where t = time since switch S was closed, i = current in the circuit, q = charge on the capacitor, L = inductance of the inductor, R = resistance of the resistor, C = capacitance of the capacitor, and $V(t)$ = voltage produced by ARB, the arbitrary waveform generator. The voltage waveform we desire to produce across resistor R is

$$V_R(t) = \begin{cases} 0, & t < 0, \\ V_0 \sin(\omega_0 t), & \tau > t > 0, \\ 0, & t > \tau. \end{cases} \tag{2}$$

Here, $\omega_0 = 2\pi f_0$ is the angular frequency (f_0 is the ordinary frequency) of the sinusoid, and τ is the "pulse length," i.e., the total time duration of the pulse, and is assumed to correspond to an integral number of half-cycles of period $1/f_0$. Mathematically, we assume $\tau = n/2f_0$, where n is a positive whole number. The quantity V_0 is an arbitrarily-selectable voltage amplitude. We choose to solve Eq. (1) subject to the realistic initial conditions that $q = 0$ and $i = 0$ at time $t = 0$.

If the voltage $V_R(t)$ of Eq. (2) is actually produced across R, the current in the circuit must be

$$i(t) = V_R(t)/R, \tag{3}$$

and the charge on the capacitor at any time t must be

$$q(t) = \int_0^t i(t)dt$$

$$= \begin{cases} 0, & t < 0, \\ (V_0/R\omega_0)[1 - \cos(\omega_0 t)], & \tau > t > 0, \\ Q, & t > \tau. \end{cases} \tag{4}$$

where Q is the (constant) charge on the capacitor at the moment of termination of the driving voltage. The value of Q depends upon whether an even or an odd number of half-cycles of a gated sinusoid is desired to appear across R. We determine Q by imposing the condition of continuity of electric charge on the capacitor from time $t = \tau^-$ to time $t = \tau^+$. The value of Q at $t = \tau^-$ is determined by using the expression for $q(t)$ in the time interval $\tau > t > 0$, given by the middle expression of Eq. (4). This means that if an even number of half-cycles is desired then $Q = 0$ [since $1 - \cos(\omega_0 \tau) = 0$ for an even number of half-cycles]. If an odd number of half-cycles is desired then $Q = 2V_0/R\omega_0$.

[since $1 - \cos(\omega_0 \tau) = 2$ for an odd number of half-cycles]. Thus,

$$Q = \begin{cases} 0, & \text{(for an even number of driving half-cycles)} \\ 2V_0/R\omega_0, & \text{(for an odd number of driving half-cycles).} \end{cases} \tag{5}$$

We can now determine the required transient-suppressing driving voltage $V(t)$ by simply substituting Eqs. (2)–(5) into Eq. (1). The result is

$$V(t) = \begin{cases} 0, & t < 0 \\ \dfrac{V_0}{R}\left(L\omega_0 - \dfrac{1}{\omega_0 C}\right)\cos(\omega_0 t) \\ \quad + V_0 \sin(\omega_0 t) + \dfrac{V_0}{\omega_0 RC}, & \tau \geq t \geq 0 \\ Q/C, & t > \tau. \end{cases} \tag{6}$$

That Eq. (6) does indeed represent the exact solution of the transient suppression problem for the LCR circuit can be justified on the basis of the uniqueness of the solution of Eq. (1). The argument is as follows: Since the solution of Eq. (1) is unique, there necessarily can be only one solution that satisfies both this differential equation and the given initial conditions. If we turn the problem around, we can assume that Eq. (6) is a *given* drive that is to be substituted for the inhomogeneous term $V(t)$ in Eq. (1). The problem then becomes that of solving Eq. (1) for $q(t)$ subject to the assumed conditions that $q = 0$ and $dq/dt = 0$ at $t = 0$. Since Eq. (6) was constructed by directly satisfying both the initial conditions and the differential equation, Eq. (4) must obviously be the unique solution of Eq. (1) for $q(t)$ given the inhomogeneous driving term represented by Eq. (6). It follows that $V_R(t)$ of Eq. (2) is the voltage that appears across resistor R in response to the drive of Eq. (6) and the given initial conditions.

For convenience, we re-write the transient-suppressing drive $V(t)$ of Eq. (6) in the form

$$V(t) = A \sin(\omega_0 t + \phi) + V_{dc}, \quad \tau \geq t \geq 0, \tag{7}$$

where

$$A = (V_0/R)\sqrt{R^2 + (L\omega_0 - 1/\omega_0 C)^2},$$

$$\phi = \sin^{-1}\left(\frac{L\omega_0 - 1/(\omega_0 C)}{\sqrt{R^2 + (L\omega_0 - 1/\omega_0 C)^2}}\right),$$

and

$$V_{dc} = V_0/\omega_0 RC.$$

Note that Eq. (7) applies only to the time interval $\tau \geq t \geq 0$, the other time intervals being covered by the first and last expressions of Eq. (6). Also note that in the form of Eq. (7), $V(t)$ is easily seen to consist of a sum of a "pedestal voltage" V_{dc} plus a sinusoidal voltage of amplitude A that is phase-shifted by an amount ϕ with respect to the sinusoid appearing across R [note Eq. (2)]. The expression for ϕ given above shows that at the circuit's undamped resonance frequency (where $L\omega_0 = 1/\omega_0 C$), $\phi = 0$. For driving frequencies above the undamped resonance (where $L\omega_0 > 1/\omega_0 C$), ϕ is positive and below the undamped resonance (where $L\omega_0 < 1/\omega_0 C$), ϕ is negative.

C. Discussion

We now consider, with hindsight, why the simple driving waveform of Eq. (7) might (possibly) have been anticipated. In what follows, it is assumed the drive frequency is sufficiently close to the undamped resonance frequency that the phase angle ϕ is negligibly small. The discussion is also restricted to the suppression of only the turnon transient, although an interpretation of the turnoff transient would be similar in nature.

If either a pedestal or gated sinusoidal drive voltage were separately applied to the LCR circuit, an exponentially damped sinusoidal transient voltage would appear across R. It is easy to show that in both the pedestal drive and gated sinusoid drive cases the exponential taper is of the mathematical form $e^{-Rt/2L}$. The frequency of the damped sinusoidal component in both cases is the resonance frequency of the circuit. Thus, it is (perhaps) not surprising that it is possible to choose a pedestal voltage amplitude V_{dc} such that the decaying sinusoid it produces will just precisely cancel the transient portion of the response to the gated sinusoidal drive, thereby entirely eliminating the turnon transient. In mathematical terms, a gated sinusoidal drive will produce a current in the circuit having the mathematical form $A_1 e^{-Rt/2L}\sin(\omega_r t) + A_2 \sin(\omega_0 t)$, where ω_r is the resonant angular frequency $= 2\pi f_r$, ω_0 is the driving angular frequency, and A_1 and A_2 are amplitudes. (This form of the current is only exactly valid when ω_0 is exactly equal to the undamped resonant angular frequency $= 1/\sqrt{LC}$.) A pedestal voltage drive will produce a current in the circuit having the mathematical form $A_3 e^{-Rt/2L}\sin(\omega_r t)$. We have seen that matters can be arranged such that $A_1 = -A_3$, so that the sum of the two responses exactly cancels the transient-producing exponential terms. The driving voltage waveform of Eq. (7) does exactly this.

Perhaps it *is* surprising that the *sign* of the applied pedestal voltage V_{dc} turns out to be the same as the sign of the amplitude of the driving sine wave A. (The sign of both A and V_{dc} are determined by the sign of V_0.) One might suppose that to cancel the transient effects produced by an initially positive-going sine wave drive one would require the application of a negative-going pedestal. However, this is not so. Since the transient behavior exhibited by the response of the circuit to a simple gated sinusoidal drive is that the initial observed amplitude of oscillation is *below* the desired steady-state amplitude, one must apply an *even-greater* voltage transient near the start of the signal in order to compensate for the deficiency in amplitude. In the case of an LCR circuit the pedestal voltage V_{dc} having the same sign as the amplitude A of the driving sinusoid gives precisely the required increase in voltage transient.

II. TRANSDUCER EQUIVALENT CIRCUIT

Although the LCR circuit considered in Sec. I captures the essence of a damped oscillating system, it is not directly applicable to the transient suppression problem for two rea-

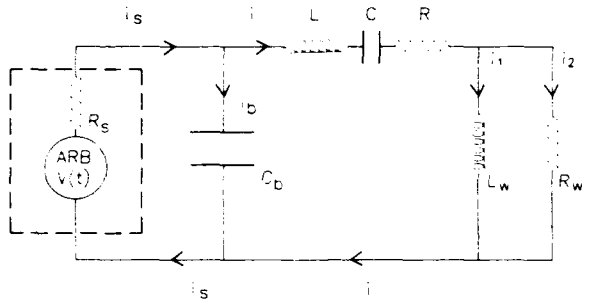

FIG. 2. Equivalent circuit of a spherical transducer driven by an arbitrary waveform generator. ARB—arbitrary waveform generator that produces voltage waveform $V(t)$, R_s—internal resistance of voltage source; C_b—blocked capacitance (includes cable capacitance); L—inductance of inductor contained in LCR branch, i.e., the motional inductance; C—capacitance of capacitor in LCR branch, i.e., the motional capacitance; R—resistance of resistor in LCR branch, i.e., the motional resistance; L_w—inductance of the reactive component of the radiation load of the fluid on the transducer; R_w—resistance of the resistive component of the radiation load of the fluid on the transducer; i, i_1, i_2, i_b, i_s—currents.

sons. First, the LCR circuit does not include the so-called "blocked capacitance" C_b of a transducer, and second, it does not account for the radiation loading of the transducer. In order to address these issues, we consider next the equivalent circuit of Fig. 2. (See, for example, Fig. 3 of Ref. 7.)

The blocked capacitance C_b of Fig. 2 should be understood to include the additional capacitance of the cable attached to the transducer. The two other additional circuit elements, denoted here by L_w and R_w, represent the effects of radiation loading on the transducer produced by projection of sound into the surrounding fluid medium. It is shown in Ref. 7 that these circuit elements represent exactly the effects of fluid loading on a spherically shaped transducer.

The element R_s is taken here to account for the departure of the voltage source from the ideal. That is, R_s represents the internal resistance of the voltage source, including the ARB and any amplifier used to drive the transducer of interest. Of course, a frequency-dependent internal impedance Z_s would be more realistic. However, as will be seen in the discussion that follows, in order for the transient suppression method to be successful, the internal impedance of the voltage source must necessarily be negligibly small. The internal resistance R_s is included simply to demonstrate the consequences of failing to use a voltage source of very low internal impedance. Including a frequency-dependent voltage source internal impedance would not materially affect the conclusions.

A. Transient-suppressing drive

We now proceed to derive the transient-suppressing drive $V(t)$ for the circuit of Fig. 2 in a manner similar to that used in evaluating the LCR circuit. Unlike the analysis of the LCR circuit, we will find that the analysis in the present case results in an approximate solution. We will therefore go through the circuit calculations in some detail so that the basis of the transient-suppressing drive is clear. (It is important to point out that the use of different Kirchhoff loops from those presented can result in a drive voltage that fails to solve the transient-suppression problem.) Also, we choose to require the voltage across the fluid-load resistor R_w to be a gated sinusoid, rather than to require this condition on the resistor R as was done in analyzing the LCR circuit. This choice is based on the assumption that the pressure waveform produced in the surrounding fluid medium will be a scaled replica of the voltage waveform appearing across the resistor R_w.

We require the voltage $V_w(t)$ across resistor R_w to be

$$V_w(t) = \begin{cases} 0, & t<0, \\ V_0 \sin(\omega_0 t), & \tau \geq t \geq 0, \\ 0, & t>\tau. \end{cases} \tag{8}$$

Here, τ again represents the pulse length and V_0 denotes an arbitrarily selectable voltage amplitude.

Since inductor L_w is wired in parallel with resistor R_w, we have the condition that

$$L_w \frac{di_1}{dt} = V_0 \sin(\omega_0 t), \quad \tau \geq t \geq 0, \tag{9}$$

which can be integrated to give

$$i_1 = \frac{V_0}{L_w \omega_0}[1-\cos(\omega_0 t)], \quad \tau \geq t \geq 0. \tag{10}$$

In producing Eq. (10) from Eq. (9), the realistic condition that $i_1 = 0$ at $t=0$ has been imposed. Similarly, conservation of charge gives

$$i = i_1 + i_2, \tag{11}$$

and, combining Ohm's law with Eqs. (8), (10), and (11) produces

$$i = \frac{V_0}{L_w \omega_0}[1-\cos(\omega_0 t)] + \frac{V_0}{R_w}\sin(\omega_0 t). \tag{12}$$

We may next apply Kirchhoff's law to the loop starting at the inductor L and proceeding through circuit elements in the order $L \to C \to R \to R_w \to C_b \to L$, thus obtaining the equation

$$iR + L\frac{di}{dt} + \frac{q}{C} + i_2 R_w = \frac{q_b}{C_b}, \tag{13}$$

where q is the charge on capacitor C and q_b is the charge on capacitor C_b. Equation (13) may now be differentiated with respect to time, and Eq. (12) and Ohm's law may be used to eliminate i and i_2 yielding

$$\begin{aligned} i_b = {} & C_b R\left(\frac{V_0}{L_w}\sin(\omega_0 t) + \frac{V_0\omega_0}{R_w}\cos(\omega_0 t)\right) \\ & + LC_b\left(\frac{V_0\omega_0}{L_w}\cos(\omega_0 t) - \frac{V_0\omega_0^2}{R_w}\sin(\omega_0 t)\right) \\ & + \frac{C_b}{C}\left(\frac{V_0}{L_w\omega_0}[1-\cos(\omega_0 t)] + \frac{V_0}{R_w}\sin(\omega_0 t)\right) \\ & + R_w C_b\left(\frac{V_0\omega_0}{R_w}\cos(\omega_0 t)\right). \end{aligned} \tag{14}$$

Once again, conservation of charge requires

$$i_s = i + i_b. \tag{15}$$

Combining Eqs. (12), (14), and (15) produces

$$i_s = \frac{V_0}{L_u\omega_0}[1-\cos(\omega_0 t)] + \frac{V_0}{R_u}\sin(\omega_0 t) + C_b R\left(\frac{V_0}{L_u}\sin(\omega_0 t) + \frac{V_0\omega_0}{R_u}\cos(\omega_0 t)\right)$$
$$+ LC_b\left(\frac{V_0\omega_0}{L_u}\cos(\omega_0 t) - \frac{V_0\omega_0^2}{R_u}\sin(\omega_0 t)\right) + \frac{C_b}{C}\left(\frac{V_0}{L_u\omega_0}[1-\cos(\omega_0 t)] + \frac{V_0}{R_u}\sin(\omega_0 t)\right)$$
$$+ C_b V_0\omega_0\cos(\omega_0 t). \tag{16}$$

Equation (16) gives the source current i_s when the transducer is driven in the transient-suppressed mode.

Finally, we write a Kirchhoff loop starting at the ARB, and proceeding around the loop defined by the circuit elements ARB$\rightarrow R_s \rightarrow L \rightarrow C \rightarrow R \rightarrow R_u \rightarrow$ARB to give

$$V(t) = i_s R_s + L\frac{di}{dt} + \frac{q}{C} + iR + V_0\sin(\omega_0 t), \quad \tau \geqslant t \geqslant 0. \tag{17}$$

The only quantity not yet defined in Eq. (17) is q, the charge on capacitor C. We obtain this by performing the integral

$$q = \int_0^t i\,dt, \tag{18}$$

where i is given by Eq. (12) and it is assumed that $q = 0$ at $t = 0$, thus producing

$$q = \frac{V_0 t}{L_u\omega_0} - \frac{V_0}{L\omega_0^2}\sin(\omega_0 t)$$
$$+ \frac{V_0}{R_u\omega_0}[1-\cos(\omega_0 t)], \quad \tau \geqslant t \geqslant 0. \tag{19}$$

Equation (17) constitutes the desired transient-suppressing drive for the turnon interval $\tau \geqslant t \geqslant 0$. Note that Eqs. (12), (16), and (19) define the quantities i, i_s, and q, respectively, which are required to evaluate Eq. (17).

In order to deduce the required functional form of $V(t)$ for the interval $t > \tau$, i.e., the turnoff interval, we will impose the requirement that the current i_1 in the inductor L_u remain constant at the value it has at the moment $t = \tau^-$. We choose to impose this condition because it assures zero voltage drop across both the inductor L_u and the resistor R_u for times $t > \tau$.

The required form of the function $V(t)$ after time $t = \tau$ depends upon whether an even or an odd number of half-cycles is desired to appear in the radiated signal during the interval $\tau \geqslant t \geqslant 0$, since the current appearing in inductor L_u at time $t = \tau^-$ depends upon this as well. From Eq. (10), we have that

$$i_1(\tau^-) = \begin{cases} 0, & \text{if } \tau \text{ corresponds to an even number of half-cycles,} \\ 2V_0/L_u\omega_0, & \text{if } \tau \text{ corresponds to an odd number of half-cycles.} \end{cases} \tag{20}$$

It follows that $i_2 = 0$ for $t > \tau$, and thus conservation of charge gives $i = i_1$ for $t > \tau$. Writing further loop equations, we find that $i_b = (C_b/C)i_1$ and $i_s = (1 + C_b/C)i_1$ for $t > \tau$. Thus, writing a loop equation for the element path ARB$\rightarrow R_s \rightarrow L \rightarrow C \rightarrow R \rightarrow L_u \rightarrow$ARB produces the result

$$V(t) = \left(1 + \frac{C_b}{C}\right)i_1 R_s + i_1 R + \frac{q(t)}{C}, \quad t > \tau. \tag{21}$$

We obtain an expression for the quantity $q(t)$ required in Eq. (21) by integrating the current which, for $i(t) = i_1 =$ constant produces

$$q(t) = i_1 t + K, \quad t > \tau, \tag{22}$$

where K is a constant of integration. The constant K can be evaluated in terms of the value of $q(t)$ at $t = \tau^-$ [note Eq. (19)] thus, in combination with Eq. (21), finally yielding

$$V(t) = \begin{cases} \dfrac{q(\tau^-)}{C} \\ \left[\left(1 + \dfrac{C_b}{C}\right)R_s + R\right]\dfrac{2V_0}{L_u\omega_0} + \dfrac{1}{C}\left(\dfrac{2V_0}{L_u\omega_0}(t-\tau) + q(\tau^-)\right) \end{cases}, \quad t > \tau. \tag{23}$$

Here, the upper expression applies in the case of an even number of half-cycles of signal output, while the lower expression applies in the case of an odd number of half-cycles. The constant quantity $q(\tau^-)$ required to evaluate Eq. (23) is itself evaluated using Eq. (19).

B. Discussion

While the analysis of Sec. II A is relatively straightforward, it is only approximate. This is so because the approach presented there has not succeeded in satisfying all initial conditions that must be imposed to evaluate the response of the circuit of Fig. 2. At time $t = 0$, these conditions are (i) $i_1 = 0$, (ii) $i = 0$, (iii) $q = 0$, and (iv) $q_b = 0$. These initial conditions are forced by the assumptions that the capacitors are initially uncharged, there are initially no currents in the circuit, and the fact that the presence of resistors in the circuit inhibits instantaneous changes in the charges and currents.

Conditions (i), (ii), and (iii) are readily seen to be satis-

fied by examining Eqs. (10), (12), and (19), respectively. However, no provision for satisfying condition (iv) has been made in the foregoing analysis. Thus, the solution for the desired transient suppressing drive, given by Eq. (17), can only be approximate. It is therefore necessary to justify the de-emphasis of condition (iv) with respect to conditions (i)–(iii), and to investigate how well Eq. (17) actually does in achieving the desired transient suppression.[5]

We justify emphasizing condition (i) on the basis of the fact that we are most interested in the observation of sound actually produced in the surrounding fluid medium. If the current i_a were not zero at time $t = 0$, a noncausal radiation would be produced by the model, a condition that is clearly unsatisfactory.

We justify emphasizing conditions (ii) and (iii) on the basis of the expectation that the LCR branch of the circuit of Fig. 2 will primarily control the transients produced in the circuit. Since we are interested in transient suppression, it is important to account for all electrical influences occurring in the portion of the circuit that is expected to be most significant in producing these transients.

We justify de-emphasizing condition (iv) on the basis of the fact that if the voltage source internal resistance R_s were zero, condition (iv) would not even be a requirement. That is, a capacitor *can* be charged instantaneously if there is no resistor in series with it. Thus, if a voltage source of very low internal resistance is used in realizing the circuit of Fig. 2, no great difficulty should arise in de-emphasizing condition (iv). In connection with the present discussion, a "low" internal resistance means that the time constant $R_s C_p$ should be much smaller than the period of the sinusoidal waveform which is desired to be produced across resistor R_a. If the frequency of this desired waveform is denoted by f_0, the condition of requiring a voltage source of small internal resistance can be expressed in the form

$$R_s \ll 1/f_0 C_p. \tag{24}$$

It should also be understood that in addition to conditions (i)–(iv), four more conditions are required at $t = \tau$. These conditions impose the requirement that the charges and currents do not change instantaneously in response to the instantaneous change in drive voltage that occurs at $t = \tau$.

C. Circuit differential equation and "derived" initial conditions

In order to address the problem of verifying that the approximate driving voltage waveform specified by Eqs. (17) and (23) is effective in actually suppressing the transient response of the circuit, we will now consider what must be done in order to obtain the exact solution of the differential equation of the circuit. This exact solution can then be used to assess the effectiveness of the proposed driving voltage waveform by directly computing the voltage that would be produced across resistor R_a if the driving voltage waveforms of Eqs. (17) and (23) were actually applied by the ARB. However, the differential equation in the present case will be seen to be sufficiently complex that we will limit ourselves to displaying the form of the equation. The solution of the equation subject to the initial conditions described above was actually carried out using a symbol manipulation computer program, and the rather involved symbolic solution will not be displayed here.

By applying appropriate loop equations, it can be shown that the current i_1 of the circuit of Fig. 2 obeys the differential equation

$$\begin{aligned}\frac{R_s C_p L L_a}{R_a}\frac{d^4 i_1}{dt^4} &+ \left[\frac{L L_a}{R_a} + R_s C_p\left(L + \frac{R L_a}{R_a} + L_a\right)\right]\frac{d^3 i_1}{dt^3} \\ &+ \left(L + \frac{R L_a}{R_a} + L_a + R_s\frac{L_a}{R_a} + R_s C_p R + \frac{R_s C_p L_a}{C R_a}\right) \\ &\times \frac{d^2 i_1}{dt^2} + \left(R + \frac{L_a}{C R_a} + R_s + \frac{R_s C_p}{C}\right)\frac{di_1}{dt} \\ &+ \frac{1}{C} i_1 = \frac{dV(t)}{dt}.\end{aligned} \tag{25}$$

It should not be surprising that the differential equation is of fourth order, in view of the fact that there are four initial conditions to be satisfied. Note also that in the limit as $R_s \to 0$, Eq. (25) becomes a third-order differential equation. This means that in this limit only three initial conditions must be satisfied. This helps to further substantiate the deemphasizing of condition (iv) in deducing the transient-suppressing drive.

The initial conditions (i)–(iv) considered above are "fundamental" conditions in the sense that they follow from the known properties of electrical circuit elements and reasonable assumptions concerning charges and currents prior to driving the circuit. Of course, before these conditions can be applied to the problem of solving Eq. (25), "derived" initial conditions must be deduced. These derived conditions must express the initial conditions (i)–(iv), as well as the additional conditions at $t = \tau$, in terms of the equation variable i_1. The required derivations follow from straightforward applications of Kirchhoff's laws.

III. NUMERICAL CALCULATIONS

In order to determine the effectiveness of the transient-suppressing driving voltage waveform proposed here, the exact symbolic solution of the differential equation of Eq. (25), subject to suitably expressed forms of initial conditions (i)–(iv) (plus the additional conditions at $t = \tau$), and with its right-hand side evaluated using Eqs. (17) and (23), was computed. These somewhat formidable calculations were carried through with the aid of the symbolic mathematics computer program SMP[14]. To allow rapid numerical evaluation of the resulting symbolic expressions, the SMP facility for generating FORTRAN subroutines from symbolic expressions was used.

As an example calculation, the USRD F56 standard transducer[9] was considered. This particular source was chosen for study since it is spherical, and thus its behavior is expected to be well described by the equivalent circuit of Fig. 2 for frequencies up to its lowest resonance frequency. This resonance frequency occurs at about 12 kHz. This transducer was also chosen because approximate values for the electrical elements of the equivalent circuit are known.[10]

(The value used here for the blocked capacitance C_b also includes the transducer's cable capacitance.) Since the internal resistance R_s of the voltage source is selectable independently of the equivalent circuit's other electrical element values, we will investigate the circuit response for a few different values of this parameter.

In Fig. 3 is presented a graph of the voltage across resistor R_u that appears in response to an ordinary gated sinusoidal driving voltage. This driving sinusoidal waveform is characterized by a frequency of 12 kHz, and is applied for a duration of three cycles. In Fig. 3, as in several subsequent figures, the "time window" used corresponds to 6 cycles of the driving frequency of 12 kHz. Hence, the response shown in Fig. 3 represents the turnon transient for all 3 cycles for which the driving voltage waveform is applied and 3 cycles of the turnoff transient. As can be seen, the response does not quite achieve steady state by the third cycle, and the turnoff transient has not yet completely died away by the end of the time window that is depicted. It should be noted that the reasonable behavior observed in Fig. 3 suggests that the equivalent network, its assumed element values, the differential equation and initial conditions used in its solution, are all quite realistic.

Next, we use Eqs. (17) and (23) to deduce the required transient-suppressing drive $V(t)$ for this particular transducer model. Since the desired 3-cycle drive contains 6 half-cycles (an even number), the turnoff portion of the drive is given by the upper expression of Eq. (23). The resulting transient-suppressing driving waveform that must be applied to the circuit by the ARB, for the case in which voltage source internal resistance $R_s = 1$ ohm, is depicted in Fig. 4. (In Fig. 4, as well as in several subsequent figures, note that the zero of time has been slightly displaced from the vertical axis, in order to more clearly exhibit the behavior of the depicted function for times near $t = 0$.) The resulting voltage waveform appearing across the resistor R_u is depicted in Fig. 5. As can be seen, the transient-suppressing drive has succeeded in eliminating virtually all of the transient behavior that is evident in Fig. 3.

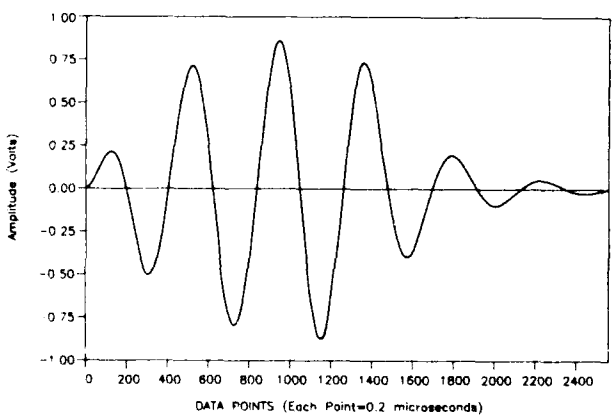

FIG. 3. Voltage waveform appearing across the fluid-load resistor R_u in the circuit of Fig. 2 in response to driving voltage waveform consisting of 3 cycles of a 12-kHz tone burst.

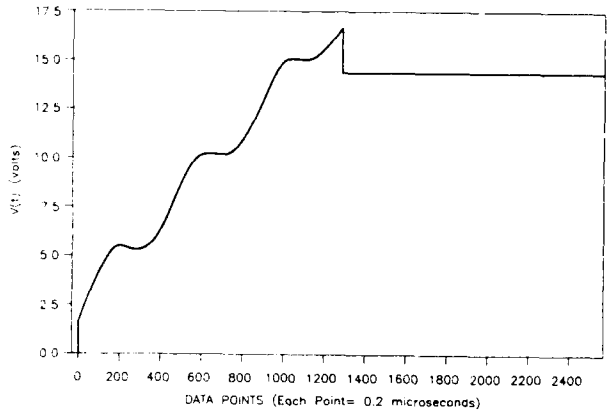

FIG. 4. Transient-suppressing driving voltage waveform that must be produced by the ARB in order to create a 3-cycle, 12-kHz tone burst across the fluid load resistor R_u in the circuit of Fig. 2: $R_s = 1$ ohm.

We investigate the result of increasing the voltage source internal resistance R_s in Fig. 6. In Fig. 6(a) is depicted the voltage across resistor R_u when the voltage source internal resistance $R_s = 10$ ohm, and Fig. 6(b) depicts the voltage across resistor R_u when the voltage source internal resistance $R_s = 100$ ohm. [In each case, the transient-suppressing driving waveform $V(t)$ has been re-computed to properly account for each new value of R_s.] The increase in transient behavior with increasing voltage source internal resistance is evident by comparing Figs. 5, 6(a), and 6(b). However, even the case for which $R_s = 100\ \Omega$ exhibits substantially suppressed transient behavior as compared to the gated-sine response depicted in Fig. 3.

The results presented in Figs. 5 and 6 can also be used to "tighten" the restriction represented by the expression of (24), assuming that the response of other transducers will be similar to that seen in these figures. If the behavior seen in Fig. 6(a) is deemed to be an acceptable transient suppression, but that of Fig. 6(b) is not, we can modify (24) to be instead

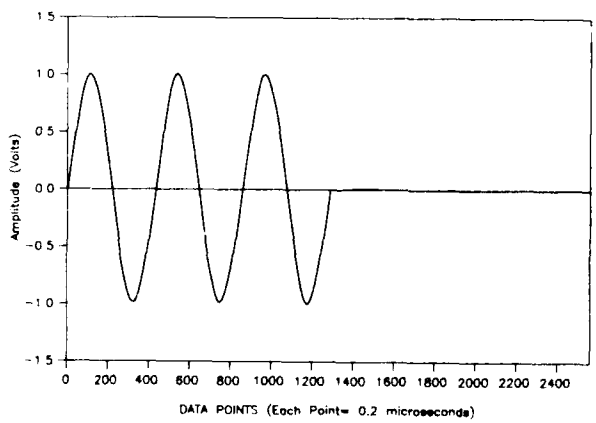

FIG. 5. Voltage waveform appearing across resistor R_u in response to the transient-suppressing drive depicted in Fig. 4: $R_s = 1$ ohm.

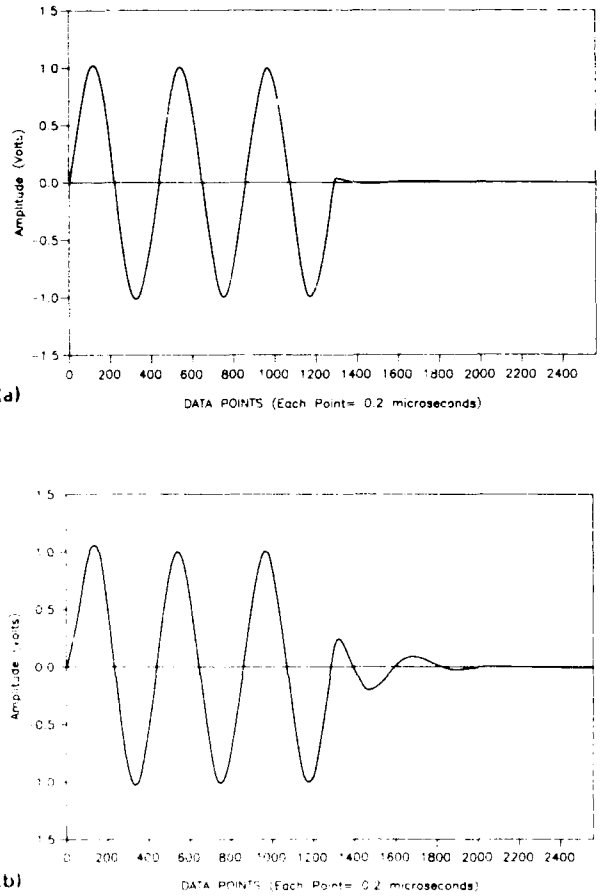

FIG. 6. Voltage waveform appearing across resistor R_a in the circuit of Fig. 2 in response to the appropriate transient-suppressing drive required to produce 3 cycles of a 12-kHz steady-state (one burst) for various internal resistances of the voltage source: (a) $R_s = 10$ ohm; (b) $R_s = 100$ ohm.

$$f_0 R_s C_b \approx 10^{-3}. \tag{26}$$

The expression of (26) has been produced by rounding to the nearest order of magnitude the value of the product $f_0 R_s C_b$ used to generate Fig. 6(a). Note that expression (26) can be used to help select an appropriate power amplifier to drive a given transducer of known blocked capacitance C_b in the transient-suppressed mode. That is, (26) is an empirical relation that sets an upper bound on R_s.

It is also interesting to examine the various currents in the circuit when the circuit is driven in the transient-suppressed mode. In Fig. 7(a)–(d) are depicted the currents i_s, i_b, i_c and i_L, respectively, for the case of $R_s = 1$-ohm source resistance. [The large current spikes seen at the beginning and ending of Fig. 7(a) and (b) represent rapid charging and discharging, respectively, of the blocked capacitor C_b.]

In view of the differing functional forms assumed by the turnoff portion of the transient-suppressing drive $V(t)$ for the even-half-cycle and odd-half-cycle cases, it is also interesting to investigate one case in which the transducer model is driven to produce an odd number of half cycles across the resistor R_a. In Fig. 8 the resulting voltage across resistor R_a is shown for the case in which it is desired to produce 5 half-cycles of a 12-kHz signal across R_a. ($R_s = 1$ ohm here.)

Before leaving the subject of the time-domain response of the circuit of Fig. 2, it is worthwhile to point out that the circuit responds to the transient-suppressing drive at frequencies away from resonance in a manner similar to that described here for resonance. Of course, in actual practice it is not expected that the responses for frequencies above resonance will be as good as those predicted by the theory, since for such frequencies the equivalent circuit is generally inapplicable (due to the influences of higher-mode resonances).

IV. THE TRANSIENT-SUPPRESSING DRIVE AND ITS PROPERTIES

Here, we put the turnon portion of the transient-suppressing drive of Eq. (17) into a slightly different form, so that its mathematical structure is more transparent. In particular, it is convenient to rewrite Eq. (17) in the form

$$V(t) = A\sin(\omega_0 t + \phi) + M_{\text{ramp}} t + V_{\text{dc}}, \quad \tau > t > 0, \tag{27}$$

where

$$A = (a^2 + b^2)^{1/2}, \tag{28}$$

$$a = \frac{-RV_0}{L_a\omega_0} - \frac{R_s V_0}{L_a\omega_0} - \frac{C_b R_s V_0}{CL_a\omega_0} - \frac{V_0}{CR_a\omega_0} + C_b R_s V_0\omega_0 + \frac{C_b LR_s V_0\omega_0}{L_a} + \frac{LV_0\omega_0}{R_a} + \frac{C_b RR_s V_0\omega_0}{R_a}, \tag{29}$$

$$b = V_0 + \frac{LV_0}{L_a} + \frac{C_b RR_s V_0}{L_a} + \frac{RV_0}{R_a} + \frac{R_s V_0}{R_a} + \frac{C_b R_s V_0}{CR_a} - \frac{V_0}{CL_a\omega_0^2} - \frac{C_b LR_s V_0\omega_0^2}{R_a}, \tag{30}$$

$$M_{\text{ramp}} = \frac{V_0}{CL_a\omega_0}, \tag{31}$$

$$V_{\text{dc}} = \frac{RV_0}{L_a\omega_0} + \frac{R_s V_0}{L_a\omega_0} + \frac{C_b R_s V_0}{CL_a\omega_0} + \frac{V_0}{CR_a\omega_0}, \tag{32}$$

and

$$\phi = \tan^{-1}(a/b). \tag{33}$$

In the form of Eq. (27), it is clear that the transient-suppressing drive for the turnon interval $\tau > t > 0$ consists of a sum of a ramp voltage, a pedestal voltage, and a sinusoidal voltage which is phase shifted by an amount ϕ with respect to the sinusoidal voltage that appears across the resistor R_a.

We next examine how each of the transient-suppressing parameters that determine the shape of the driving waveform of Eq. (27) vary with respect to frequency. In particular, we examine how the parameters A, V_{dc}, M_{ramp}, and ϕ vary as the frequency of the signal that is desired to be radiated into the fluid that surrounds the transducer is changed. For the purpose of these calculations, we again use the transducer parameters from Ref. 10.

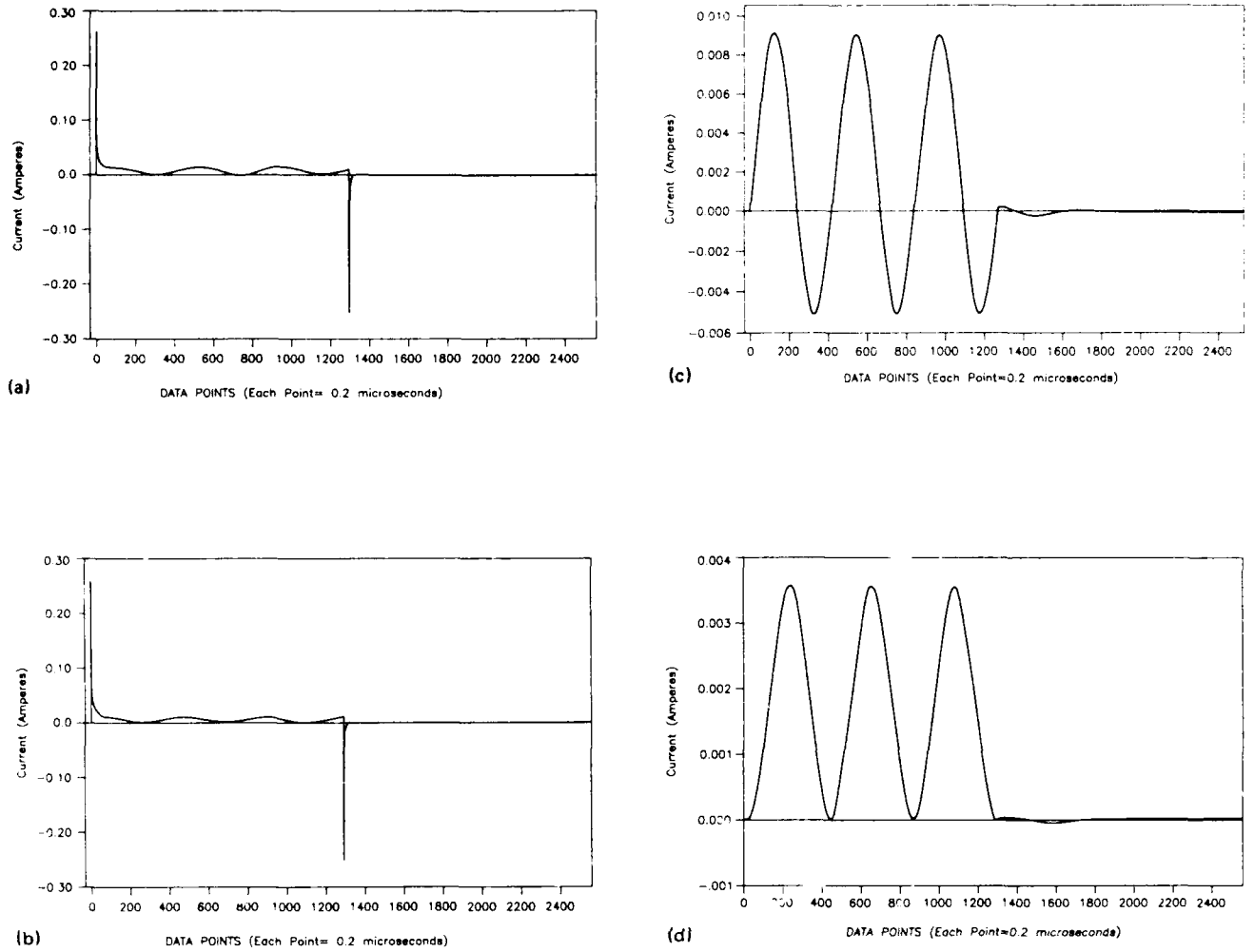

FIG. 7. Various currents in the circuit of Fig. 2 when the ARB produces the appropriate transient-suppressing driving waveform required to produce a 3-cycle, 12-kHz tone burst for internal resistance R = 1 ohm: (a) i; (b) i_n; (c) i; (d) i_1.

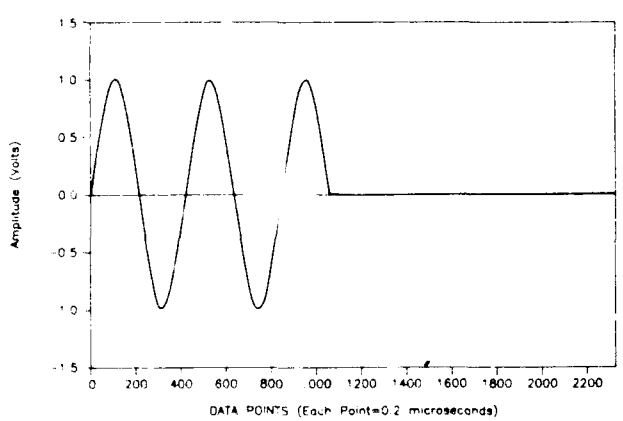

FIG. 8. Voltage waveform appearing across resistor R in the circuit of Fig. 2 in response to the transient-suppressing drive required to produce a 5/2-cycle, 12-kHz steady-state tone burst for R = 1 ohm.

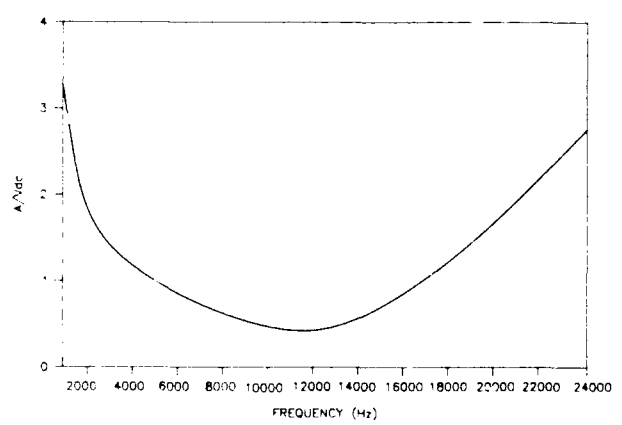

FIG. 9. Variation of the ratio of the amplitude of the sinusoidal component to the pedestal voltage component of the transient-suppressing drive A/V_1 as a function of the frequency of the tone burst desired to appear across resistor R. Horizontal axis begins at 1000 Hz.

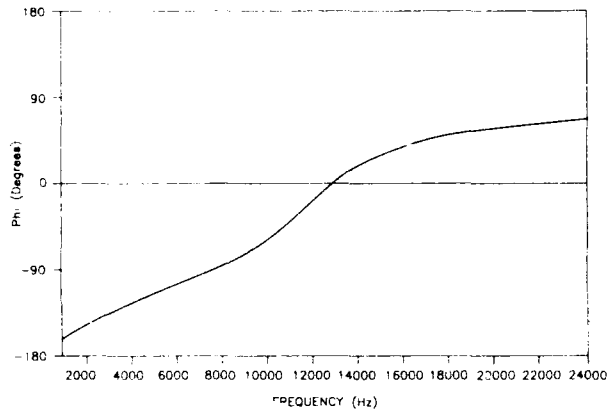

FIG. 10. Relative phase ϕ between sinusoidal component of the transient-suppressing driving voltage waveform and radiated waveform as a function of frequency. Horizontal axis begins at 1000 Hz.

In Fig. 9 the variation in the ratio A/V_{dc} with respect to frequency is depicted. As can be seen, this ratio is less than 1 in the vicinity of the (approximately) 12 kHz resonance frequency of the F56 transducer, but is greater than 1 for frequencies that are either significantly greater, or significantly less, than the resonance frequency. This means that the pedestal voltage V_{dc} must be greater than the amplitude of the sinusoidal amplitude A near resonance, but is less than the sinusoidal amplitude for frequencies far from resonance.

The phase angle ϕ between the sinusoidal portion of the transient-suppressing waveform and the waveform appearing across resistor R_a (i.e., the waveform radiated into the surrounding fluid medium) is plotted as a function of frequency in Fig. 10. It is clear from Fig. 10 that the phase angle ϕ is close to zero for frequencies near resonance. However, the phase angle ϕ is less than zero for frequencies below resonance and is greater than zero for frequencies above resonance. (This behavior is qualitatively similar to the behavior of the phase angle in the case of the LCR circuit that was described in Sec. I.) Analytically evaluating Eq. (33) in the limit as frequency approaches zero shows that ϕ approaches -180 deg in this limit. Similarly, evaluating Eq. (33) in the limit as frequency approaches infinity shows that ϕ approaches $+180$ deg in this limit. (In the case of the LCR circuit, ϕ varies from -90 deg to $+90$ deg as frequency varies from zero to infinity.[11]) These limiting values are consistent with the behavior of the function ϕ that is apparent in Fig. 10. (This surprising behavior of the phase angle ϕ, viz., that the sinusoidal component of the driving waveform approaches the condition of being 180 deg out of phase with respect to the radiated waveform, will be considered further in connection with the discussion of Fig. 12.)

In Fig. 11 the variation in the slope of the ramp voltage M_{ramp} with respect to frequency is depicted. This function varies hyperbolically with frequency, as is evident from the reciprocal dependence of the quantity M_{ramp} upon angular frequency ω_0 seen in Eq. (31).

It is clear from the rather large values of M_{ramp} seen in Fig. 11 that the total duration τ of a transient-suppressed waveform must be significantly restricted. For example, assuming a maximum allowable peak transmitting voltage of 1000 V, the maximum value of τ at 12 kHz is 0.0173 s. This corresponds to about 200 cycles of 12-kHz sound. While 200 cycles may not seem to be a great pulse-length restriction, this calculation assumes only a 1 V drop across the resistor R_a. This is only 0.1% of the maximum amplitude of the driving waveform of 1000 V under consideration here. If it is desired to produce a 10-V drop across resistor R_a, which is still only 1% of the total drive voltage amplitude, the maximum pulse length is reduced to only about 20 cycles. Since a very low voltage source internal resistance is required for the transient-suppression method to work, as described in Sec. II, the 1000-V maximum source amplitude discussed here might be difficult to achieve in practice. For a maximum voltage source amplitude of 100 V, and a desired voltage drop of 10 V across resistor R_a, the maximum achieveable pulse length would, of course, reduce to just 2 cycles. (As is described in a companion publication,[12] actual experiments

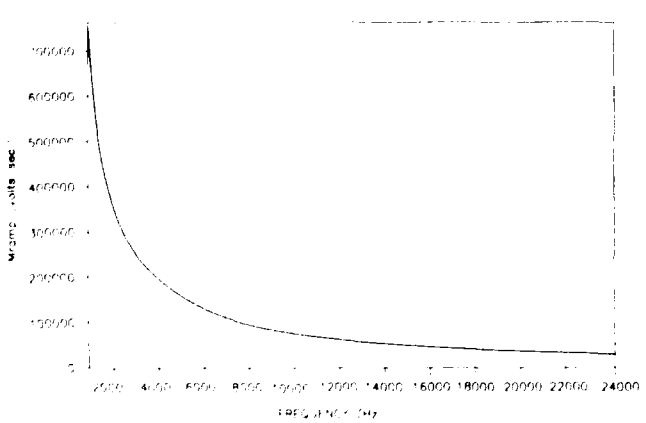

FIG. 11. Slope of the ramp voltage component of the transient-suppressing driving voltage waveform M_{ramp} as a function of frequency. Horizontal axis begins at 1000 Hz.

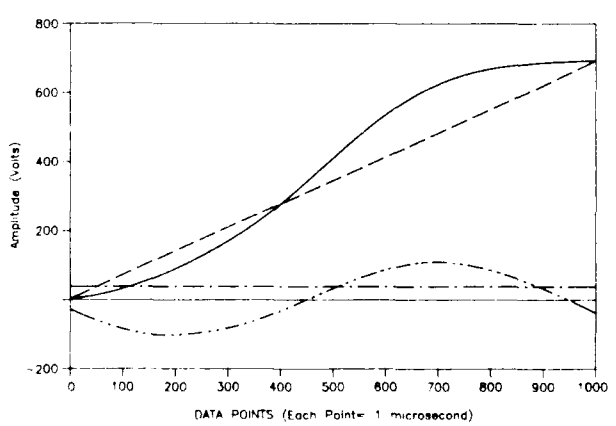

FIG. 12. Components of the transient-suppressing drive: ---- pedestal voltage V_{dc}; — sinusoidal voltage $A\sin(\omega_0 t+\phi)$; ---- ramp voltage $M_{ramp}t$; total transient-suppressing drive $A\sin(\omega_0 t+\phi)+M_{ramp}t+V_{dc}$. Frequency of tone burst desired to appear across resistor R_a is taken to be 1 kHz here. Horizontal solid line is zero-voltage reference.

have shown that achieving pulse lengths of up to several cycles is quite realizable in practice.)

Finally, in Fig. 12 the time-domain representation of each of the three separate components of the transient-suppressing drive is shown together with the total transient-suppressing drive, i.e., the sum of the three components. In producing Fig. 12, the driving frequency was taken to be 1 kHz, and a 1-cycle portion of the drive is shown. Also, the sinusoid appearing across resistor R_u is taken to be initially positive going. We examine this case because of the surprising result, mentioned above, concerning the zero-frequency, and infinite frequency, limiting behaviors of the transient-suppressing drive. In particular, it is certainly surprising that a 180-deg phase shift, relative to the signal that is desired in the fluid, is required in the sinusoidal component of the transient-suppressing drive. What this means, of course, is that far from resonance, if an initially positive-going sinusoid is desired to appear across resistor R_u, then an initially negative-going sinusoidal component of the transient-suppressing drive is required!

As can be seen by examining Fig. 12, the overall transient-suppressing drive (curved solid line) actually departs smoothly from zero in this case, and in the positive-going direction, despite the fact that the sinusoidal component (double-dot dash line) is indeed initially negative going. It is clear from Fig. 12 that this smooth positive-going behavior of the transient-suppressing drive is achieved by the influence of the positive-going ramp voltage; i.e., the positive contribution of the ramp voltage (dashed line) clearly compensates the negative contribution of the negative-going sine. The pedestal voltage contribution (dot-dash line) essentially cancels the nonzero value of the negative-going sine at times near $t = 0$. (The negative-going sine is nonzero at time $t = 0$ because, at a frequency of 1 kHz, the required phase angle ϕ is approximately -162.5 deg, not -180 deg as it is for the limiting zero-frequency drive.)

V. SUMMARY, CONCLUSIONS, AND DISCUSSION OF FUTURE WORK

It has been demonstrated that it is possible to apply a suitably shaped driving voltage waveform to a spherical transducer equivalent circuit and produce a good approximation of a steady-state tone burst voltage waveform across the portion of the circuit that represents radiation loading. The suitably shaped driving-voltage waveform, here termed the transient-suppressing drive, consists of a sum of a pedestal voltage, a ramp voltage, and a sinusoidal voltage which is phase shifted with respect to the sinusoid appearing across the radiation-load circuit elements. One disadvantage of the transient-suppressing drive is that it exhibits very low efficiency, from the point of view of the percentage of the total applied voltage available to produce sound radiation. This percentage is typically less than 10%, and can be less than 1%. While the low radiation efficiency obtained using the transient-suppressing drive is unfortunate, it does not preclude the utility of the method, as demonstrated in the experiments described in a companion article.[12]

Although the case of a spherical transducer was of particular interest here, the method presented is also applicable, at least in an approximate way, to more complicated transducers. For example, the successful application of the method to a transducer array is described in Ref. 12.

Future work will consider extensions of the method to other transducer types. These include flexural disks, moving coils, tonpilzs, and various arrays of transducer elements. Two approaches will be considered. The first, which is largely empirical, will involve attempting to apply the spherical transducer equivalent circuit of Fig. 2 to nonspherical transducer types. The second approach will involve using equivalent circuits that are more suitable to the transducer types of interest. However, the calculation of the transient-suppressing drive will follow the same methods of analysis presented here.

It is expected that the present method will have an immediate application to the second of the two areas of interest described in the Introduction, namely, scattering or reflection measurements conducted in a confined region. Application to the first area described, i.e., transducer calibration in a confined region, is currently restricted to calibration of spherical sources only. Whether application to sources of other type is feasible must await the results of the further research discussed here.

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