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BROAD-BAND TRANSMITTER SYSTEM DESIGN

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1. INTRODUCTION

The term "bandwidth" has a variety of interpretations. Sometimes it refers to the frequency band over which the power output of a transducer is within a specified range, usually 3dB; at other times the limits refer to other parameters such as the projector sensitivity or source level of the system. These definitions generally assume that sufficient power is available from the amplifier to maintain a constant drive voltage (or current) over the frequency band of interest. When amplifier power is limited, however, the variation in admittance of the transducer over this band may make this assumption questionable, and it is then preferable to take the possible variation in applied voltage or current into account and to base the definition of bandwidth on the overall performance of the system.

This approach is adopted in this paper, and the transducer is considered as an element in the total system, the bandwidth being interpreted in terms of how effectively power is transferred from the electrical source to be radiated at the output as acoustic power. The system is analysed as a filter, to give results indicating optimum parameters for the system. The design optimisation is more readily achievable with Tonpilz designs than with other low frequency sources.

2. ANALYSIS AS A FILTER

The behaviour of the electrical admittance of a piezoelectric transducer around resonance can be represented adequately by a simple equivalent circuit as shown in Fig 1. If an inductance L_0 is connected in parallel with the electrical capacitance C_0 to resonate with it at the resonance frequency of the transducer, then the circuit resembles a half-section band-pass filter and its behaviour may be analysed by standard filter methods. For this, it is useful to define various parameters:-

Bandwidth factor $$W^2 = C_1/C_0 = k^2/(1 - k^2) \quad (1)$$

where k is the coupling coefficient of the transducer, ie where $k^2 = C_1/(C_0+C_1)$

Frequency variable $$y = (1/W)(\omega/\omega_s - \omega_s/\omega) \equiv \Omega/W \quad (2)$$

where ω_s is the transducer's angular resonance frequency, ie $\omega_s^2 = 1/L_1 C_1$

Nominal Resistance $$R_N = 1/(\omega_s W C_0) = \omega_s L_1 W \quad (3)$$

Matching Factor $$ß = R_1/R_N = 1/(Q_M W), \text{ where } Q_M = \omega_s L_1/R_1 \quad (4)$$

In most cases, the loss resistor R_0 is large enough to be neglected for this analysis. The input conductance G_i and susceptance B_i may then be expressed in terms of these parameters by:-

$$G_f = \frac{1}{R_N} \frac{\beta}{(\beta^2+y^2)} \qquad (5)$$

$$B_f = \frac{1}{R_N} \frac{y(\beta^2+y^2-1)}{(\beta^2+y^2)} \qquad (6)$$

Normalised admittance curves obtained by plotting $\beta R_N B_t$ against $\beta R_N G_t$ are shown in Fig 2 for values of ß from 0.6 to 1.2. These curves illustrate how the shape of the curves depends on the value of the matching factor ß. The frequency varies along the curves, the value of the normalised frequency parameter y/β being indicated by the dashed vertical lines. The frequencies corresponding to $y = \pm 1$ are usually called the nominal filter cut-off frequencies, whilst the frequencies given by $y = \pm\beta$ are those at which the conductance of the transducer falls to half its value at resonance.

For those applications in which it is important to obtain maximum power from the amplifier over a wide band, the admittance values of the transducer which are presented to the amplifier as the frequency is varied within the band must be restricted to lie within some reasonable limits. When these limits have been specified, a value of the matching parameter ß can be selected which optimises the shape of the curve so that it maximises the frequency band for which the transducer's admittance falls within these specified limits. For example, if the range of loads is required to be such that its magnitude should not vary by more than 2:1, and that its phase angle should not exceed that corresponding to a power factor of 0.8, then curve (b) in Fig 2 represents the shape giving the widest bandwidth. Thus, the optimum value of ß for these limits is approximately 0.8. Expressed in other terms, this corresponds to an optimum given by $WQ_M \approx 1.25$ (or, since the coupling coefficient k is slightly less than W, $kQ_M \approx 1.2$.) The fractional bandwidth is then approximately equal to the value (k) of the coupling coefficient of the transducer [1].

This result emphasises the importance of the coupling coefficient parameter in determining the maximum bandwidth of the system. However, it is important to note that it is the *combination* of k and Q_M which determines the shape of the admittance loop, and that the maximum bandwidth of the system is only achieved if the transducer is designed to have the appropriate value of the motional Q-factor Q_M to match the effective coupling factor k. Too low a value for Q_M causes too large a variation in phase angle of the load, whilst too high a value of Q_M causes too large a variation in conductance. For a lead zirconate titanate transducer, with a typical coupling coefficient of 0.5, the value of Q_M should be about 2.4, a value which is rather lower than is easily achieved for some designs.

In practice, effects due to manufacturing tolerances, environmental variations, ageing of the ceramic, and acoustic interactions between elements in an array cause the load admittance to vary by considerably more than than the 2:1 range for the ideal case, an overall variation of 6:1 in magnitude and phase angles up to 60° being quite typical. Although this may seem to weaken the case for seeking an optimum value of kQ_M, the admittance variation due to frequency is in fact a major contributor to the overall range, and it is therefore generally advisable to aim at the optimum since any markedly different value would further increase the range of either modulus or phase angle. From the reliability point of view, the primary requirement of the driving amplifier should be to withstand the overall range of load admittances without causing failure, otherwise catastrophic breakdown of the system may occur.

In addition to avoiding failures of the driving amplifiers, it is necessary to consider the safety of the transducers themselves, and the control of the acoustic output. If the amplifier acted as a constant voltage source, a range of 6:1 in load admittance would imply a similar variation in input current to the transducer, and for a correctly tuned element this would correspond to about the same variation in piston velocity. Such a large deviation from the nominal piston velocity might well result in mechanical failure of the transducer. Similarly, a constant current source driving into this range of admittances could result in electrical failure in the elements, due to the excessive swings in voltage. These variations would need to be taken into account in developing the transducers by increasing the safety factors in the design, or by devising methods to control the variations in applied voltage and current. The simplest and least expensive method is to adopt the second approach, by making the output impedance of the amplifier comparable with the average load impedance. This is a well known technique for maximising the power transfer from a source into a resistive load, and it has the advantage also that neither the voltage nor the current into the load can exceed twice the nominal voltage for any positive value of the load. A safety factor of two applied to the nominal design values should then be adequate to ensure safety of the transducers, provided that the source impedance is matched to the nominal load.

Up to this point, the choice of source impedance has been considered in terms of preventing actual failure of the system, and this is of course the prime requirement. However, it is usually necessary also to meet certain acoustic requirements over the whole operating band, allowing for the effects of the variations in admittance. The system may be treated as one in which power is transmitted from the source through the filter network to be radiated as acoustic power in the load resistor. The transfer characteristics of such a filter depend on its terminating resistors, and the optimum characteristics are obtained when the driving source has an output resistance of $R_N/ß$, to "match" the load resistance of $ßR_N$. Since ß is not quite equal to unity, this matching condition is not exactly the same as that above, but the differences are not in practice very significant. This matching optimises the power transfer into the nominal load, taking into account its variation in admittance over the whole band. It also reduces the variations in piston velocity which can occur because of admittance variations in admittance which are caused by acoustic interactions between the elements of an array. Carson [2] proposed the use of a high source impedance to give good "velocity control" near the resonance frequency, but this becomes less effective near the edges of the frequency band. Using a matched source gives less precise control at resonance, but provides better control over the whole band [3]. It should thus improve control of the acoustic output as well as the reliability of the system.

3. BROAD-BAND NETWORKS

Analysis of the transducer as a half-section band-pass filter suggests that it may be possible to achieve better characteristics by adding further elements to the filter. For example, addition of a series *LC* combination to form the full section ($n=3$) network shown in Fig 3 can give further control of the response curve. Details of the design techniques for such networks have been described by Stansfield [4], and examples of the admittance loops resulting from the application of such methods to a lead zirconate titanate Tonpilz transducer are shown in Fig 4. The curve for the $n=2$ case in Fig 4a is similar to that for the parallel tuned transducer considered above. Addition of further arms in the network adds further loops in the admittance curves. The dotted circles represent the admittance loci corresponding to an output power 0.13dB less than the nominal value; this is the reduction appropriate to an overall 2:1 range of admittance when driven by a matched source. For this example, for which the coupling coefficient is assumed to have a value of 0.48, the optimal Q_M values and the resulting fractional bandwidths are:-

For n	= 2	3
Q_M	= 2.6	2.15
Fractional bandwidth	= 0.38	0.60

An increase of some 50% in the calculated bandwidth should thus be feasible by this method. A small further increase could be achieved by adding additional parallel and series *LC* combinations, although the effects diminish in significance as extra sections are added.

These networks can be realised by adding electrical components to the piezoelectric transducer. It is worth considering whether they could be realised instead by mechanical elements within the transducer itself. However, the equivalent circuits used in this paper are based on the impedance system of analogues, in which there is no mechanical equivalent to the parallel inductor L_0 with one terminal connected to ground, and this places a restriction on what networks are mechanically realisable. The "coupled resonator" network shown in Fig 5 can be a good approximation to the full network over a narrow band, and can be realised by mechanical elements since it does not involve any inductors connected to ground. A Tonpilz design modified to correspond to such a network would require the addition of a suitably tuned spring and mass connected in front of the piston. The calculated bandwidth for an example of such a design is 0.53, an improvement over the n=2 case above, but not so good as for the n=3 example using the full network.

4. LOW FREQUENCY SOURCES

The optimising methods above have generally been applied to Tonpilz designs, for which a reasonable degree of independent control of the coupling coefficient and motional Q-factor is possible, and it is therefore often feasible to satisfy the $kQ_M \approx 1.2$ criterion well enough. For other low frequency designs, such as the flexural disc or flextensional transducers, it is less easy to control the relevant parameters, because of their interdependence amongst themselves and with others. Expressions for the characteristics of thin flexural disc transducers have been given by Woollett [5]. These lead to the expression for Q_M in terms of the radius *a* and thickness *t* as:-

$$Q_M = 7.02(1 + 0.103a/t)^{3/2}\eta_{ma} \qquad (7)$$

For the idealised case where the motional-acoustic efficiency η_{ma} is taken as 100%, the value of Q_M is a function only of a/t, and ranges from over 100 for $a/t=50$ to 13 for $a/t=5$. Woollett quotes the theoretical coupling coefficient for a lead zirconate titanate flexural disc transducer as 0.4, so the optimum valus of Q_M should be about 3. Achieving this value would need an exceptionally thick disc, which would take it beyond the range of applicability of Woollett's equations and would also increase the resonance frequency markedly. It thus appears very difficult to design a flexural disc transducer to achieve the optimum condition: in practice, the coupling coefficient is often lower than the value quoted by Woollett, so that the matching value of Q_M is greater than 3, and it may well be easier to achieve this higher optimum value.

Design relationships for the motional Q-factor of flextensional transducers are not well established. Brigham and Glass [6] stated that the widest bandwidth would be achieved by using a long, thin, highly eccentric ring, but recognised that this would be limited by practical engineering considerations. Oswin and Dunn [7] gave an empirical relationship for Aluminium Shell Class IV designs as

$$3adf^2Q_M = 1$$

where f = frequency in kHz, a = semi-major axis, and d = overall height in mm. For GRP shells, the numerical factor in this equation is 10 instead of 3. For practical designs, Oswin quotes typical values of Q_M as being between 2 and 6, with values of coupling coefficient between 0.25 and 0.35. It is therefore possible that a typical Class IV flextensional transducer may have a value of kQ_M which is near the optimum, although there appear to be few systematic design rules available to control the parameters of the transducer in order deliberately to attain this optimum.

5. REFERENCES

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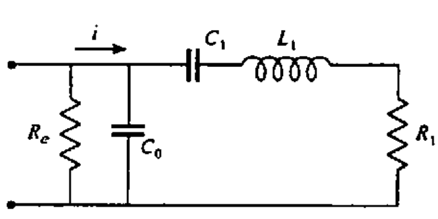


Fig 1. Electrical equivalent circuit of Piezoelectric transducer near resonance.

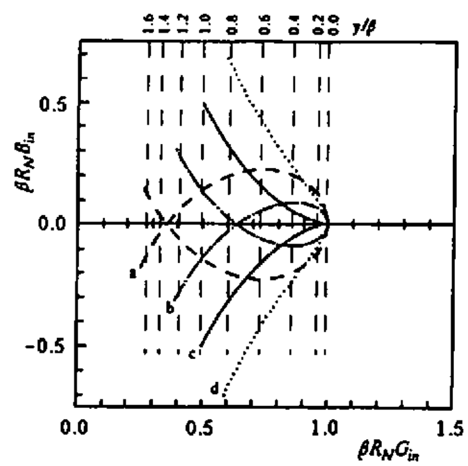


Fig 2. Admittance curves for band-pass filter
(a) ß = 0.6 (b) ß = 0.8
(c) ß = 1.0 (d) ß = 1.2

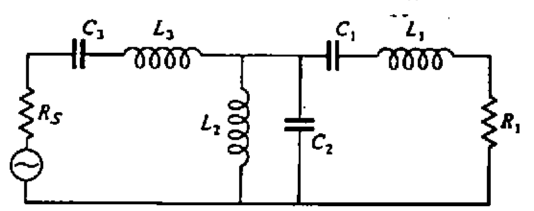
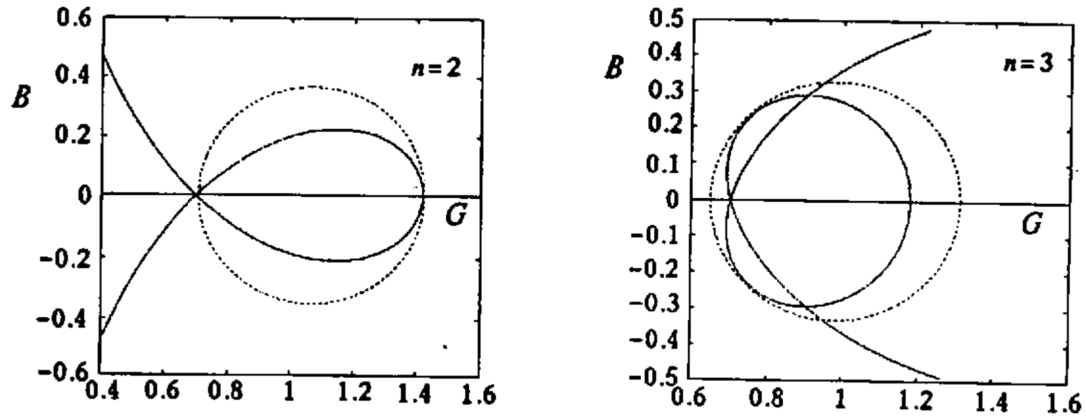


Fig 3. Band-pass filter network, (n = 3)



(a) (b)

Fig 4. Admittance curves for optimised band-pass filter networks: (a) n=2; (b) n=3.

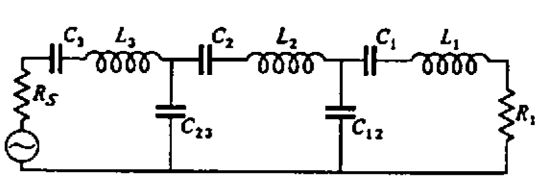


Fig 5. Coupled resonator network.