ELECTRICAL MATCHING OF UNIDIRECTIONAL SURFACE WAVE DEVICES

Roy B. Brown
Texas Instruments Incorporated
Dallas, Texas 75222

Abstract

Recent advances in surface wave device (SWD) technology have made it possible to obtain bandpass filters with insertion losses as low as 0.65 dB by implementing 3-phase unidirectional transducers. This paper discusses the techniques used for electrically matching unidirectional SWD's and shows recent experimental results obtained by using these techniques. Also included are discussions on electrical measurements of 3-phase transducer terminals, different methods of phase splitting and transducer interface, component value calculations for phase splitting and matching, and a review of sources of filter insertion loss.

I. Introduction

Recent reports[1,2] have discussed principles and development of low loss surface wave device bandpass filters using three-phase unidirectional transducers[3]. One of the major problems in implementing unidirectional surface wave devices (USWD's) has been the additional interface and matching complexities caused by using 3-phase transducers rather than the usual single-ended 2-phase transducers. This paper discusses the different techniques for matching to 3-phase transducers, in addition to presenting some recent experimental results using these techniques.

One of the major difficulties in using 3-phase transducers is that the interactions of the different terminals of the transducer with each other prevent simple single-port electrical characterization of the terminals. Taking either 3-port S-parameter measurements or 2-port measurements (with the third port grounded) effectively circumvents this problem. Also, if great care in measurement and interpretation is made, a simple single-ended impedance measurement can be made by grounding one port and measuring the other two ports in parallel and converting the results to equivalent S-parameters. After obtaining the S-parameter measurements (or equivalent), an S-matrix can be derived, which properly characterizes the 3-phase transducer under any external electrical condition. It should also be noted here that the above measurements can also be equivalently obtained for unidirectional transducers of 4 or more ports, but the added complexity in most cases does not justify utilizing transducers of more than 3 ports.

II. Transducer Interface Methods

Thus far, three methods of interfacing 3-phase transducers to a single-ended source have been found. The first method, shown in Figure 1, avoids using any phase-shift network. The phase shifting from 2- to 3-phase and 3- to 2-phase is accomplished by cascading two USWD's as shown in the figure. The only external electrical matching required is that of the center transducers, which are matched the same as any conventional transducer. In addition, a single inductor between each of the three ports of each unidirectional transducer is required for optimum matching of the complex transducer impedances. Even though the interface component complexity is significantly reduced by using the above circuit, several disadvantages remain. One of the main disadvantages is the necessity of using two devices which doubles insertion loss and size. Another important disadvantage is that this particular method encourages circulating acoustic waves and therefore decreased triple-transit suppression unless the center transducers perfectly absorb all incident acoustic energy. Wideband experimental devices have been fabricated using the above interface methods, which had approximately 8 dB insertion loss and 1.5 dB ripple at the center frequency, but increasing ripple away from the center frequency.

The second method of interface, shown in Figure 2, has been used previously with good results[1,2], particularly for wideband matching. The circuit consists of matching networks from each of the 3 transducer ports to a 3-port phase splitter. The 3-phase splitter can use either a lumped-component delay circuit or transmission line of appropriate length and impedance. The primary disadvantage of this second method of interface is the excessive number of components required which discourages its use in practical applications.

The third method of interface, reported previously in theory[3], but only recently implemented in practice is shown in Figure 3. With this method, only a single 60° phase shifter and one matching network is needed to provide the correct phase relationship between the terminals if one of the terminals is grounded. Figure 4 demonstrates by vector relationships how this is possible. The phase shifting for this circuit must be done directly between unmatched terminals, because any matching between the ungrounded terminals and the 60° phase shifter will introduce unsymmetric phase shifts between the transducer terminals. Incorporating a pi-network as the 60° phase shifter gives a resultant interface circuit with a minimum number of components, thereby making it easiest for "tweaking" to optimum performance. However, using a pi-network may produce a somewhat narrower interface bandwidth than the 3-phase interface method discussed earlier.

III. Interface Circuit Analysis

The problems of implementing transducer interface circuitry, using the last two methods of interface discussed above, can be reduced to two major areas: designing an adequate phase shifter and implementing the device S-parameter measurements in such a way to determine required circuit component values. Phase shifting an arbitrary angle can be done by three methods used thus far: distributed transmission lines of appropriate length and impedance, a pi-network, and a bridged-T network. A distributed transmission line would find application at high frequencies (>300 MHz) where lumped-component circuits would be impractical to implement because of excessive parasitics.

A pi-network[4] is one of the simplest "lumped element" circuits that can provide a desired phase shift at a given frequency. Assuming a series inductor (L) and equal value parallel capacitors (C) and an equal input and output impedance (R_o), the circuit component values at frequency ($\omega=2\pi f$) and phase lag (θ) are:

$$L = \frac{R_o \sin\theta}{\omega} \qquad C = \frac{R_o \sin\theta}{\omega(1-\cos\theta)} \qquad (1)$$

Using a pi-network rather than the equivalent T-network allows even further simplication for the 60° phase shift interface circuit because some of the components conveniently combine together with the matching network components.

A bridged-T circuit[5], shown in Figure 5, is more complex than a pi-network but allows a greater bandwidth in matching. In fact, the circuit impedance can be designed to remain constant over any frequency range provided the following relationships between circuit components are maintained: $L_1=4L_2$ and $L_2=R_o^2\ (C/2)$, where R_o is the circuit impedance. The phase shift at frequency ($\omega=2\pi f$) is given by the following derived formula:

$$\theta = \tan^{-1}\left[\frac{2(\omega L_2+1/\omega^2C^2X)\ L_1/XC+R_o^2-(L_1/XC)^2}{2\ R_o(\omega L_2+1/\omega^2C^2X)\ -\ 2\ L_1R_o/XC}\right]$$

where $X=\omega L_1-\ 2/\omega C$. The best way to use this formula for determining component values at a given angle and frequency is by graphical solution. It has been found that the phase change as a function of frequency for the bridged-T circuit is only slightly greater than that for an equivalent transmission line, thereby making this network very useful for wideband phase shifting.

The procedure for matching a three-port transducer using the 3-phase splitter described earlier can be summarized as follows: the 3-port S-matrix of the transducer is converted to the equivalent Y- or Z-matrix (as needed) and then modified by the addition of trial values of matching network components. The modified matrix is then converted back to S-parameters and checked for optimum matching to the 3-phase splitter. If matching is not optimum, then the trial values of the matching network are modified accordingly until after several iterations, the S-matrix is matched.

Interfacing the SWD transducer with a single 60° phase shifter required a somewhat different procedure. With this method, 2-port S-matrix measurements are taken and then converted to the equivalent Y-matrix. Given this Y-matrix and the 60° phase lag constraint between the two ports, a single-ended admittance can be derived at each of the two ports. Using the value of admittance for port-2 (see Figure 3), and assuming $\theta=60°$, the pi-network component values can be derived by using equation (1). Combining the components from the pi-network with the Y-matrix gives a modified Y-matrix from which a new admittance for port-1 can be derived. The derived port-1 input admittance can then be matched as a single-ended impedance with simple 2-element matching networks[6].

IV. Experimental Results and Sources of Loss

The last method of interfacing using a single 60° phase shifter looked the most promising in terms of practical implementation; therefore experimental results will be shown only for that particular method. Figure 6 shows the most noteworthy result thus far obtained — a surface wave device with an insertion loss of only 0.65 dB, the lowest loss known to be reported thus far. This device had a center frequency of 34 MHz and used 5% bandwidth unweighted transducers on a lithium niobate substrate. A relatively low center frequency was initially chosen for these experimental devices to avoid additional fabrication and performance problems appearing at higher frequencies. Figure 6 also shows the very low level of periodic triple-transit ripple which is inherently obtainable with unidirectional transducers. The "impulse ripple" seen in the figure is most likely due to resonant plate modes which can easily be suppressed in future devices. Figure 7 shows the sidelobe response of the same device which can certainly be optimized with suitable weighting techniques[1], and also by improved plate mode suppression techniques. Figure 8 shows the response of a wider band (20% bandwidth) device of the same frequency also with unweighted transducers on lithium niobate. Even though the insertion loss of this device is somewhat greater (2.5 dB) in comparison to the narrower bandwidth device, the sidelobe suppression has considerably improved. This improvement is due to the inherent matching network "rolloff" because of the wide transducer bandwidth required and also because of the higher electrical Q of the wideband transducers. Figure 9 shows the actual interface networks with nominal circuit component values as used for both of the above experimental devices. These circuit component values were derived using the procedures described earlier and required little "tweaking" to obtain the above experimental results.

In order to obtain the very low insertion losses described above, a study was made of different sources of filter loss. Some of the sources of loss include: reflection mismatch loss, resistive parasitic losses in the phase shifting and matching networks components, improper amplitudes and relative phases between transducer terminals, parasitic resistance losses in the transducers, and diffraction and beam spreading losses. Any of the above losses can contribute significantly (>0.5 dB) to overall loss if care in design is not taken. Although approximate quantitative estimates have been made for some of the above sources of loss, the best procedure is to minimize all losses as much as possible. Reflection mismatch loss and improper amplitude and relative phase between transducer terminals can be minimized by careful optimization of shift and matching network component values. Using high-Q phase-shifting and matching components minimizes component parasitic effects, which are especially significant for wideband transducers. Parasitic resistance losses in the transducer can be minimized by using as thick of aluminum fingers as practical (usually 5,000 to 10,000 Å) and minimizing beamwidths without causing excessive diffraction. Diffraction and beamspreading losses are minimized by placing the transducers as close together as possible without causing excessive cross talk and also by making the beamwidth as wide as possible without causing excessive parasitic resistive losses. Other sources of loss that are usually negligible (<0.1 dB) include: phase errors caused by diffraction or velocity changes under the fingers, end effects in transducers due to unequal driving strength between middle and end fingers , energy absorption due to different plate modes in the substrate, and propagation attenuation on the substrate surface.

V. Conclusion

For future USWD's, even further external component reduction is possible. For example, at any design bandwidth, if the unidirectional parameters are adjusted such that the equivalent parallel resistance of Y (as shown in Figure 9) is 150 ohms, only 3 components are required for matching and phase shifting. Furthermore, if the transducer is designed to have an electrical Q of approximately 1.7 (which corresponds approximately to 8% bandwidth on lithium niobate substrates) as well as having the above parallel resistance of 150Ω, only two electrical matching and phase shifting components are required, which is the same number as usually used for matching conventional transducers.

In conclusion, it has been shown that very low loss surface wave device filters using 3-phase unidirectional transducers can be fabricated and matched with insertion losses as low as 0.65 dB. The ultimate insertion loss obtainable for future devices will depend on the bandwidth required and the type of substrate material used. Nevertheless, for any bandwidth the unidirectional transducer clearly gives an insertion loss reduction of 6 dB and almost complete triple-transit suppression as compared to conventional bi-phase transducers. Applications for filters with low losses of the magnitude now obtainable include receiver front end filters, transmitter filters, or any other application where low noise figure, high efficiency, or large dynamic range is required. A systematic transducer interface and matching procedure has been derived showing how 3-phase transducers can be interfaced with a minimum number of components, thereby allowing practical implementation of USWD filters for many of the applications mentioned above.

Acknowledgements

The author wishes to thank C.S. Hartmann and R. C. Rosenfeld for helpful discussions and insights and also L. A. Simonson and T.L. Shoquist for fabrication and assembly of the experimental devices.

References

1. R.C. Rosenfeld, R.B. Brown, & C.S. Hartmann, "Unidirectional Acoustic Surface Wave Filters with 2 dB Insertion Loss," 1974 IEEE Ultrasonics Symposium Proceedings, Cat. No. 74 CHO 896-1SU.
2. R. C. Rosenfeld, C.S. Hartmann, & R.B. Brown, "Low Loss Unidirectional Acoustic Surface Wave Filters," Proceedings of 28th Annual Symposium on Frequency Control," U.S. Army Electronics Command, Fort Monmouth, New Jersey (1974).
3. C.S. Hartmann, W.S. Jones, and H.G. Vollers, "Wideband Unidirectional Surface Wave Transducers," IEEE Transactions - Sonics and Ultrasonics, SU-19 (July 1972), pp. 378-381.
4. F.E. Terman, Radio Engineers' Handbook. New York: McGraw-Hill Company, Incorporated (1943), pp. 210-213.
5. P.R. Geffe, Simplified Modern Filter Design. New York: John F. Rider Publisher, Incorporated (1963), pp. 72-75.
6. E.N. Phillips, "Ell-Network Charts Simplify Impedance Matching," Microwaves, 7 (May 1968), pp. 44-58.

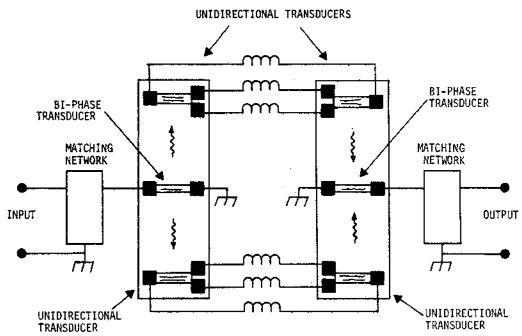


FIGURE 1. CASCADED UNIDIRECTIONAL DEVICES

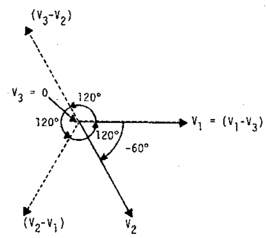


FIGURE 4. VECTOR DIAGRAM SHOWING PHASE RELATIONSHIPS BETWEEN TRANSDUCER PORTS

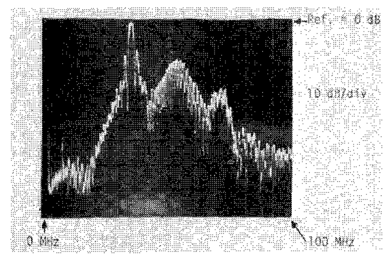


Figure 7. Narrowband USWD Sidelobe Frequency Response

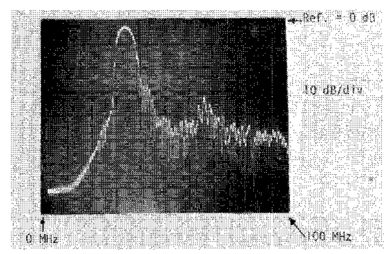


Figure 8. Wideband USWD Sidelobe Frequency Response

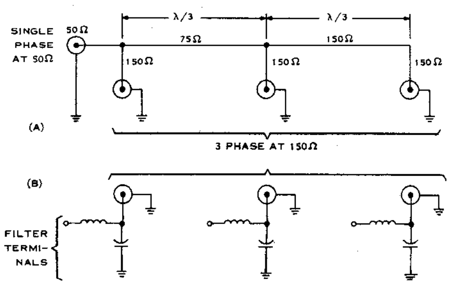


FIGURE 2. THREE-PORT TRANSDUCER INTERFACE CIRCUIT

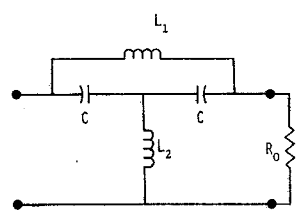
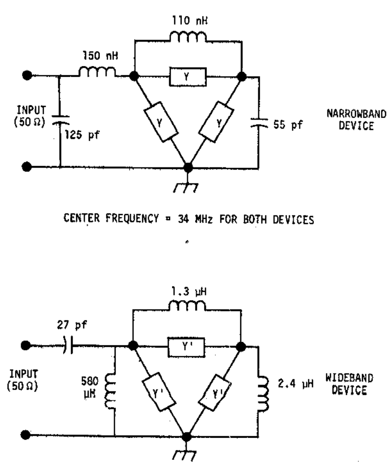


FIGURE 5. BRIDGED-T CIRCUIT



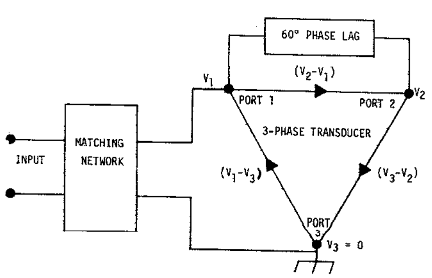


FIGURE 3. UNIDIRECTIONAL TRANSDUCER INTERFACE WITH A SINGLE 60° PHASE SHIFTER

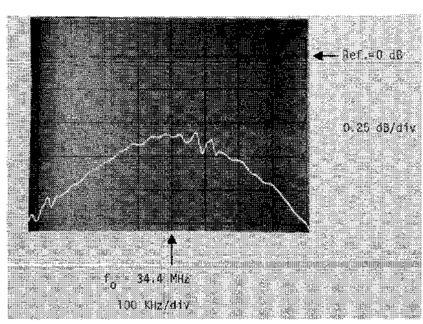


Figure 6. Narrowband USWD In-Band Frequency Response

FIGURE 9. EXPERIMENTAL NETWORKS USED FOR UNIDIRECTIONAL TRANSDUCERS