

Surface Acoustic Wave Multistrip Components and Their Applications

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Invited Paper

Abstract—The multistrip coupler is a directional coupler which operates on freely propagating surface acoustic waves with broad bandwidth and low loss. The applications of simple multistrip couplers with straight strips are discussed together with descriptions and properties of a family of multistrip components that can perform particular operations. Stepped couplers can perform the function of collinear beamwidth compression and a surface-wave “magic T” may be constructed. Bent, U-shaped, and J-shaped structures are considered that may redirect or reflect acoustic beams, and results are presented for a multistrip unidirectional transducer. Complex acoustic circuits may be built up using these elements. A beam may be reflected and stepped to the side by one track width, and both directions of a delay line may be used sequentially to double its capacity. The spurious triple transit signal of a delay line is shown to be virtually eliminated at the expense of an extra 3-dB single transit loss.

I. INTRODUCTION

THE multistrip coupler (MSC), which is simply an array of parallel metallic strips deposited on a piezoelectric substrate, can transfer acoustic power from one acoustic track to another. Up to 100-percent power transfer can be achieved with little loss in the structure. Because the energy is transferred by means of metallic strips into which bends can be incorporated, the acoustic tracks are not constrained to be parallel to each other. This flexibility has led to a family of novel devices, derived from the simpler MSC described in [1], that offer significant improvements in the performance of various surface acoustic wave devices and also perform functions not previously available to the SAW system designer.

It is the primary objective of this paper to present an appraisal of this new class of components, multistrip components, and to indicate their practical applications.

Several of the multistrip components to be presented here have previously been discussed [2]–[4], others have not, e.g., the “magic T.” In this paper we take the opportunity of subdividing multistrip components into categories derived from their basic structure. Thus we have sections devoted to the simple multistrip structure (Section II), the stepped coupler structure (Section III), and angled coupler structures (Section IV). Applications arise in which various multistrip components are combined into acoustic circuits; these are discussed in Section V.

All experimental observations reported were obtained using *Y*-cut LiNbO₃ substrates with sound propagating parallel to the *Z* axis. There are several reasons for this choice: 1) *K*² is high, which means that the number of strips required to perform a given function is not excessive; 2) multistrip components have a broad bandwidth and this can be most effec-

tively exploited on a high *K*² material; 3) the anisotropy of LiNbO₃ is such that sound propagating along the *Z* axis is effectively collimated, broadening by diffraction being only approximately one quarter of that in an isotropic material [5]. Free acoustic beams approximately 100 wavelengths wide (as generated by a 50- Ω transducer [6]) will follow closely defined tracks over appreciable distances and thus propagate as though guided. This is an important factor in the practical construction of acoustic subsystems (Section V) that are in many respects analogs of familiar microwave circuits.

The operating frequency of all the devices to be described is arranged to be below the stopband of the coupler. The simple theory of the MSC presented in [1, sec. III] is then adequate for the purpose of determining *N_T*, the number of coupler strips required for complete energy transfer. The minimum value of *N_T* is obtained when the strips are approximately as wide as gaps between them. In this case a practical formula for designing MSC's on *Y*-cut *Z*-propagating LiNbO₃ is

$$N_T = \frac{195 f/f_0}{1 - \cos(153 f/f_0)} \quad (1)$$

where *f* is the working frequency, *f₀* is the stopband frequency, and the cosine angle is in degrees. We find that a useful rule of thumb is to set the working frequency at approximately three quarters of the stopband frequency where the coupler bandwidth is near 120 percent and coupler strip reflection effects due to the proximity of the stopband are negligible.

Throughout this paper we refer to interdigital transducer parameters by the shorthand notation *n/f/l*, where *n* is the number of finger pairs, *f* is the synchronous frequency, and *l* is the aperture in wavelengths. Coupler parameters are given as *N/f₀*, where *N* is the number of separate strips in the design and *f₀* is the stopband frequency. All multistrip components had coupler strips approximately equal to the gaps between them.

II. APPLICATIONS OF THE SIMPLE MSC

A. Full-Transfer-Length Coupler

Here we consider the application of couplers with the correct number of strips to give 100-percent transfer of energy from one track to another. They are referred to as full-transfer-length couplers.

1) *Bulk Wave Suppression*: The proportion of bulk wave to surface-wave energy radiated by an interdigital transducer increases as the number of finger pairs is reduced [7]. Thus one penalty of increasing the bandwidth of an SAW delay line by reducing the number of finger pairs is that spurious bulk wave signals become significant. The inclusion of a simple full-

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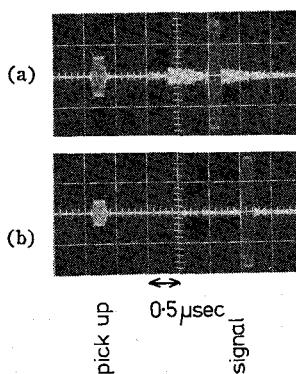


Fig. 1. (a) Output of delay line using single finger pair transducers, showing a surface-wave signal superimposed on a background due to bulk acoustic waves. (b) Output signal of delay line incorporating a full-transfer-length MSC.

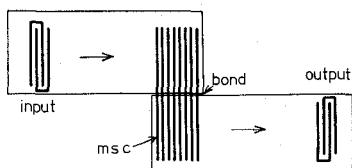


Fig. 2. Use of MSC to transfer surface waves between adjacent substrates.

transfer-length MSC into a lithium niobate delay line reduces the bulk wave signal by 15 to 20 dB while leaving the surface-wave transmission and bandwidth unaffected.

Fig. 1(a) shows the output signal at 43 MHz from a delay line consisting of a pair of single finger pair transducers of finger width 20 μm and gap 20 μm . An unacceptably large bulk wave signal is evident. Fig. 1(b) shows the improvement obtained by inserting a full-transfer-length 140/86 MSC into the delay path.

2) *Coupling Between Separate Substrates*: A full-transfer-length multistrip coupler has been used to couple surface acoustic waves from one substrate to an adjacent substrate, as shown in Fig. 2. When both substrates were lithium niobate, virtually 100-percent transfer could be achieved. Indeed, the behavior of the composite was indistinguishable from a single crystal and it would be possible to build up substrates of any size in this way. This technique could be used with the reflecting track-changer device, described in the following paragraphs, to build long delay lines. The question of coupling between dissimilar materials is raised in the discussion.

3) *Aperture Transformation*: As discussed in [1], if the tracks of the coupler are of unequal width, only a fraction of the energy is transferred in the transfer length. This energy, though, being launched from a different aperture, will have changed diffraction loss and can be intercepted by a transducer of different impedance to the launching transducer.

A particularly important situation arises when dealing with aperture weighted (finger length weighted) structures. By introducing a full-transfer-length coupler as shown in Fig. 3, the output of an aperture weighted transducer can be converted to full aperture while maintaining the original weighting. If diffraction effects are ignored, this is readily shown using [1, eq. (22)] and may be rendered obvious by working the structure in reverse (the beam from the standard transducer passes through the full-length coupler unchanged

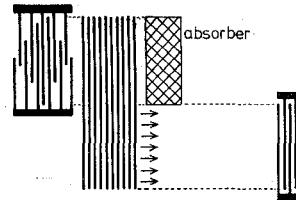


Fig. 3. Conversion of finger length weighted transducer output to constant aperture.

and is received on the weighted transducer as though fed directly). The implications are that 1) weighted transducers can be used in conjunction with the coupler, and 2) advantages can be gained by so doing since diffraction losses normally suffered by signals leaving the narrow aperture regions may be reduced if the coupler is sited close to the weighted transducer.

B. Half-Transfer-Length (3-dB) Coupler

A coupler of half the length required for complete energy transfer splits an input signal into two outputs of equal intensity, the signal in the coupled track having a 90° phase advance over the signal remaining in the original track. This phase advance is a basic property of the coupler action and as such is independent of frequency at frequencies removed from the stopband. The amplitude balance between tracks, as distinct from the phase difference, is frequency dependent. For example, it is found that an imbalance of ± 0.5 dB can only be obtained over a bandwidth of 25 percent for a working frequency of 0.75 of the stopband frequency. A modification aimed at increasing the coupler bandwidth is discussed in Section VI.

C. Fractional-Transfer-Length Coupler

1) *Precision Attenuator*: An effective means of increasing the triple transit suppression (TTS) of a simple SAW delay line is to increase the attenuation suffered by the propagating acoustic waves. If the increase in the single transit loss (STL) is β dB, then the increase in the TTS will be 2β dB. An MSC may simply be used as a precision attenuator that has the added attraction of rejecting bulk acoustic waves. The attenuation of a signal coupled by an MSC from one track to another as a function of the number of strips N is given by

$$\beta = -20 \log_{10} \sin \left[\frac{NFK^2}{2} \left(\frac{\sin^2 \theta/2}{\theta/2} \right) \right] \quad (2)$$

where $\theta = \pi \alpha f / f_0$ and the symbols are as defined in [1, sec. I].

It is calculated in [1] that spurious signals caused by reflections from the coupler itself are negligible at frequencies below, say, 90 percent of the stopband frequency. This has been confirmed in practical devices.

2) *Delay Line Tap*: Consider the simple tapped delay line of Fig. 4(a). A severe defect of this structure is that each tap reflects energy most strongly at the frequency which it is designed to receive. Also, if the application calls for the taps to be connected together each will retransmit a fraction of the power received by all of the others. The effect of a line of such transducers is to introduce a complex pattern of spurious signals. The device shown in Fig. 4(b) uses fractional-transfer-length MSC's to overcome this problem. Reflection from the multistrip tap is avoided by designing its stopband to be well removed from the operating frequency. Energy not absorbed

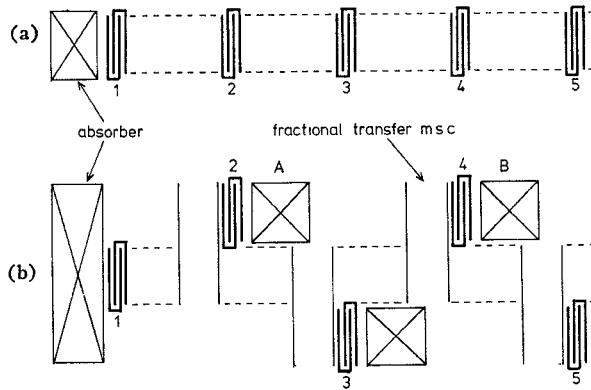


Fig. 4. (a) Simple tapped delay line. (b) Use of fractional-transfer-length couplers to isolate taps.

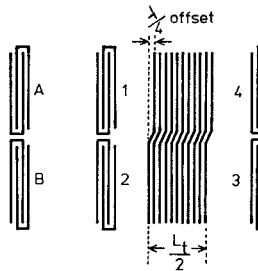


Fig. 5. MSC stepped to perform function of magic T.

by the transducer 4, for example, is removed by absorber *B*, and most of the energy reflected by 4 is absorbed by *A*. The small fraction of the reflected energy that is coupled into the delay line cannot be received by any of the other taps (because they are directional) and may be disposed of at the transmitter transducer with the aid of the echo-trap construction to be described in Section V-B.

III. COMPONENTS DERIVED FROM STEPPED COUPLER STRUCTURES

As previously stressed, it is an intrinsic feature of the MSC that the signal transferred by it from one track to another has a 90° phase advance over the signal remaining in the original track. Without violating this condition the coupler may be modified to produce arbitrary additional phase shifts in one track relative to the other. For example, the strips in one track may be displaced along the propagation direction relative to those in the adjacent track, thereby producing a stepped structure (see Fig. 5). Some interesting properties emerge from these simple structures.

A. SAW Magic T

If the step in the coupler is equal to $\frac{1}{4}$ wavelength and the length is half a transfer length, then the structure behaves as a magic T, familiar as a microwave component [8].

In this structure, a signal injected at one port will be split into two of equal magnitude which are either in phase or anti-phase with each other. Numbering the ports as in Fig. 5 and referring to *a* and *b* as input and output vectors, respectively, we have

$$b_1 = \frac{1}{\sqrt{2}} (a_3 + a_4)$$

$$\begin{aligned} b_2 &= \frac{1}{\sqrt{2}} (a_3 - a_4) \\ b_3 &= \frac{1}{\sqrt{2}} (a_1 + a_2) \\ b_4 &= \frac{1}{\sqrt{2}} (a_1 - a_2). \end{aligned} \quad (3)$$

It has the same scattering matrix,

$$|S| = \frac{1}{\sqrt{2}} \begin{vmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 0 & 0 \\ 1 & -1 & 0 & 0 \end{vmatrix} \quad (4)$$

as the magic T [8].

The device will only work as indicated at the frequency for which the offset is exactly $\lambda/4$ and at which the coupler is exactly $\frac{1}{2}$ transfer length long. The degradation in performance as the frequency varies from the design frequency may be found by using the simple theory of the coupler discussed in [1, sec. III]. It may be shown that the ratio of power between the signal leaving port 3 and that entering port 1 (excluding transducer loss) is given by

$$\begin{aligned} \frac{P_3}{P_1} &= \sin^2 \phi \cdot (1 - R^2) + R^2 - R \sin 2\phi \sin kx \\ &\quad - R \sin 2\phi \sin \pi f/2f' \end{aligned} \quad (5)$$

where $R = a_2/a_1$, $\phi = (NFK^2 \sin^2 \theta/2)/\theta$, and $\theta = \pi \alpha f/f_0$. f' is the design center frequency of the device, and other symbols are as defined in [1].

The expected performance was experimentally verified using the arrangement shown in Fig. 5. 3/61/90 transducers were positioned at each port and at *A* and *B*. The coupler stopband was at 86 MHz and the offset between strips in each track was 14 μm . The number of active strips in the coupler was adjusted by breaking strips at their midpoint with a laser burner until the coupler split a signal input at one port into two signals of exactly equal amplitude. 52 strips were found to be necessary.

Ports 1 and 2 were then driven with equal amplitude signals, in phase, and the output at port 3 was found to be 3 dB larger than a calibration signal received on either of two transducers *A* and *B* positioned to receive the backwards-directed wave from transducers 1 and 2. The signal on port 4 was more than 25 dB below this calibration signal. The phase of the signal at port 2 was then inverted and the situation at the output ports reversed.

Curve *q* of Fig. 6 shows, as a function of frequency, the decibel ratio between the output at port 3 and the output from the calibration transducer *A* when ports 1 and 2 were driven with equal amplitude signals in phase, and curve *p* shows the relative output at port 4. When the input transducers were driven antiphase, curve *s* was obtained for port 4 (identical to curve *q* within experimental error) and curve *r* for port 3. The device was also used with ports 3 and 4 as inputs (not plotted in the figure). In all cases the port carrying the sum signal gave the expected intensity gain of $3 \text{ dB} \pm \frac{1}{2} \text{ dB}$ across the measured bandwidth, but the frequency of the peak loss in the curve for the difference signal varied between 60 and 70

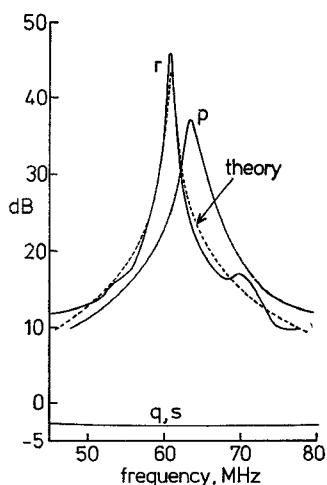


Fig. 6. Performance of magic T with equal inputs to ports 1 and 2. Inputs in phase: q , output from 3; p , output from 4. Inputs out of phase: s , output from 4; r , output from 3.

MHz. This can be accounted for by small differences in the efficiencies of the different transducers. The dashed curve in Fig. 6 was obtained using (5) with $\alpha = 0.85$ and $FK^2 = 0.043$. The fit to the results is reasonable.

For effective operation the difference signals (curves p and r of Fig. 6) should be low. This criterion dictates the useful bandwidth of the device. If we take a difference signal of more than 25 dB down as acceptable, the bandwidth is 12½ percent.

B. Beamwidth Compressor

The magic-T performance previously described could have been achieved without introducing steps in the 3-dB coupler, the required phase shift being achieved by displacement of the transducers or by introducing thin-film overlays. We discard the latter because it introduces significant technological problems and also increases the area occupied by the structure. The advantage of the stepped coupler over a displaced transducer arrangement is that it functions on a normally propagating beam; this is exploited in the beamwidth compressor.

It is obvious from inspection of (3) that a uniform wavefront ($a_1 = a_2$) incident over the complete width of the stepped coupler will only produce an output over half its width; the beam undergoes width compression by a factor of 2. By cascading these structures large acoustic power densities can be achieved for modest interdigital voltage inputs, a feature of value in nonlinear SAW devices. This approach to acoustic power intensification does not require the sophistication in design demanded by other approaches such as curved transducers or SAW lenses.

Besides acting as an intensifier, it should be noted that the beamwidth compressor can, of course, expand the width of a beam and, in fact, plays the role of an acoustic transformer.

The performance of a three-stage beamwidth compressor has been assessed using the structure shown schematically in Fig. 7. Designed for operation at 86 MHz (coupler offsets of 10 μm), the input aperture was 4.3 mm and each coupler had 57 8- μm strips and a stopband at 108 MHz. The output of the narrowest compressor was observed with a potential probe [10] and compared with the backwards-traveling beam from the input transducer, as shown in the figure. An intensity gain of 8½ dB was measured, only ½ dB below the theoretical maxi-

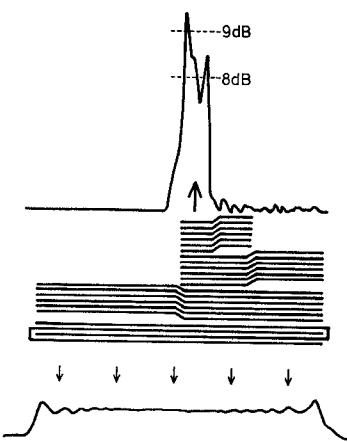


Fig. 7. Schematic of three-stage beamwidth compressor including profiles of output beams.

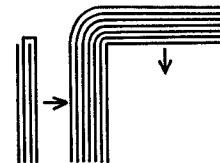


Fig. 8. Use of MSC to redirect acoustic beam.

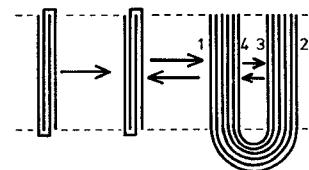


Fig. 9. U-shaped MSC as reflecting element.

mum of 9 dB. Other measurements indicated that the intensity gain was greater than 8 dB over a bandwidth of 35 percent about 86 MHz.

IV. COMPONENTS DERIVED FROM ANGLED COUPLER STRUCTURES

A. Beam-Redirection Element

An obvious derivative of the multistrip structure is the bent coupler shown in Fig. 8. If the surface-wave velocity and coupling coefficient are the same in the input and output directions, then the design is trivial provided that allowance is made for the curved connecting wires as considered in the following paragraphs. If the properties are not equivalent, the design problem is the same as that of coupling between two different materials, as discussed in [1, sec. V].

Equivalent directions could be exploited on Z -cut LiNbO_3 which has three such directions separated by 120° [9]. A series of 120° bands might be used to construct a long delay line in which the signal follows a path in the form of a series of interlaced equilateral triangles. Such structures have not been investigated experimentally.

B. Multistrip Surface-Wave Reflector

By increasing the angle between the arms of the coupler to 180° so that the coupler becomes U shaped, as shown in Fig. 9, another multistrip component is formed, a broad-band

reflector [2], which in its current form introduces only 2 dB of loss. The number of strips in the structure is chosen so that a signal input at port 1 will split into two equal signals. These leave the coupler at ports 3 and 4 in phase quadrature and reenter the opposite ports. The coupler then recombines the two signals into a single output at port 1. The device thus acts as a reflector. If acoustic power loss in the curved portion is ignored, it has the same large working bandwidth as a full-length coupler.

Design Considerations: We have chosen to use nested semicircular strips to connect together the straight active regions of the multistrip reflector because these offer the shortest connections that require no increase in the resolution of the photolithography used to fabricate the structure. The connections have several properties that affect the functioning of the reflector: 1) they introduce loss due to their electrical resistance; 2) they radiate acoustic power; and 3) they increase the number of strips necessary to form a 3-dB coupler. We consider these in turn for a standard U structure on LiNbO_3 which has a 100-wavelengths-wide acoustic aperture and the gap between ports 3 and 4 effectively zero. The results of [1, sec. V] indicate that the loss due to electrical resistance will add no more than a few tenths of a decibel to the reflection loss for 5000-Å-thick Al strips. No attempt has been made to estimate the loss due to radiation of acoustic power, but from potential probing it has been established that the total loss to surface waves is more than 13 dB down on the main reflected signal. The number of strips necessary to form a 3-dB coupler is increased because the connecting wires form a capacitive load on the active parts of the strips. If the number of strips required for a simple half-transfer-length coupler is $N_T/2$, then the analysis presented in [1, sec. V] may be used to show that the number of strips required by a U-shaped reflector is given by

$$\frac{N_T'}{2} = \frac{N_T}{2} \left(\frac{8l + 4\pi r}{8l - \pi d N_T} \right) \quad (6)$$

where d is the strip repeat distance, l is the active aperture of the input port, and r is the radius of the smallest semicircle. Techniques for modifying the reflection performance by, for example, SiO_2 underlays are considered in Section VI.

Results are presented here for a multistrip reflector with 60 U-shaped strips 8 μm wide repeated at 16- μm intervals (stopband at 107 MHz) and an active aperture of 4 mm. There was no central gap between ports 3 and 4. The surface-wave signals to be reflected by the structure were first generated by a $2\frac{1}{2}$ finger pair transducer and then passed through a simple coupler to exclude bulk waves. An untuned $2\frac{1}{2}$ finger pair transducer situated close to port 1 served to probe the signals as they approached and were reflected from the structure. Photographs of the output from this transducer at frequencies of 60, 80, and 100 MHz are given in Fig. 10(a). At each frequency there is little visible pulse distortion (even at 100 MHz, only 7 MHz from the stopband), and the reflected pulse is 2 dB smaller than the input pulse. Fig. 10(b) shows that the loss against frequency curve is flat over a bandwidth of at least 60 percent, limited only by the coupler stopband and the transducer frequency response.

C. Multistrip Unidirectional Transducer

In principle, a unidirectional transducer (UDT) may be perfectly matched into an electrical load to give zero transduction loss and zero reflections. The TTS of a delay line

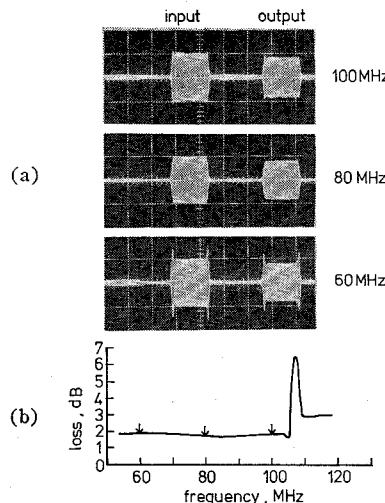


Fig. 10. (a) Signals observed at input port of U-shaped reflector. (b) Reflection loss versus frequency.

incorporating a pair of UDT's should therefore be superior to one using simple interdigital 3-port transducers which must always give rise to reflections if any attempt is made to match them into an electrical load.

The U-shaped coupler previously described may be used to combine the two oppositely directed output signals of an interdigital transducer into a single output signal [2], thus making such a transducer unidirectional. Using these structures, delay lines having 3-dB insertion loss, broad bandwidths, and low TTS have been produced.

In construction of the UDT, a transducer with an odd number of fingers is placed within the arms of a U so that the center of the central finger is offset, relative to the center line of the U, by $\frac{1}{8}$ wavelength towards the desired output port. The two signals arriving at ports 3 and 4 then differ in phase by 90° . In the case of a transducer with an even number of fingers, the required phase quadrature is achieved by offsetting the midpoint of the central gap $\frac{1}{8}$ wavelength away from the desired output port.

The same design considerations apply to the UDT as to the U-shaped reflector except that loss due to resistance and acoustic generation will be reduced because the structure is only traversed once by the acoustic wave. However, insertion of the transducer into the U means that the length of the curved portions will be increased and due allowance must be made for this in determining the number of strips. The number to be used is found from (1) and (6).

The STL and TTS of a delay line consisting of a pair of series-tuned multistrip UDT's are plotted against frequency in Fig. 11. A total minimum loss of only 3 dB was observed for this delay line. ($7\frac{1}{2}/61/108$ transducers were enclosed within $65/86$ U-shaped couplers.)

In common with other UDT designs [6], [12], [13], the TTS is large at the synchronous frequency but reaches unacceptable levels at frequencies still within the 3-dB passband of the single transit signal. The bandwidth may of course be increased by decreasing the number of finger pairs in the transducer, but we have shown previously [2] that the load impedance required to give maximum TTS is then less than 50Ω ; a $4\frac{1}{2}/107/108$ transducer required a load of only 15Ω . We have constructed a $4\frac{1}{2}$ finger pair multistrip UDT that had a maximum TTS for a 50Ω load, simply by reducing the transducer aperture. A laser burner was used to adjust the number of

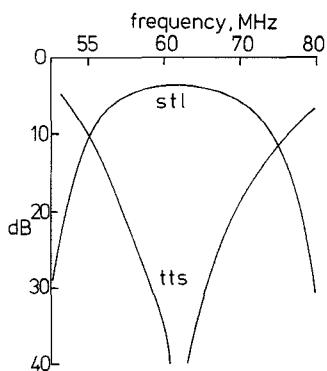


Fig. 11. TTS and STL versus frequency for a pair of multistrip UDT's.

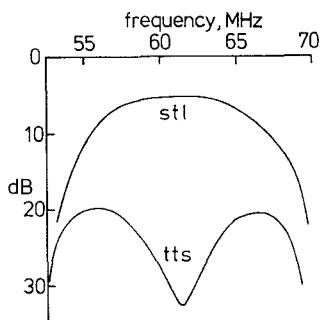


Fig. 12. TTS and STL for bidirectional transducer feeding a multistrip UDT.

finger pairs and to progressively reduce the active width of the transducers and U-shaped couplers of the 61-MHz delay line previously described. At a finger width of 4.7 mm (81 wavelengths), a 35-dB TTS was obtained over a 10-percent bandwidth for a $4\frac{1}{2}$ -dB minimum STL.

Smith *et al.* [6] have calculated, and numerous workers have confirmed, that a transducer on LiNbO_3 with an aperture of 108 wavelengths and many finger pairs has a series impedance of 50Ω . A transducer identical to that used in the 61-MHz delay line previously described was converted to $4\frac{1}{2}$ finger pairs by breaking 4 fingers with a laser burner, and a load resistance of 33Ω was then required to minimize the transducer reflection at the resonant frequency. A resistance of only 20Ω was required when the device was further modified to have only $2\frac{1}{2}$ finger pairs.

There are at least three explanations for this effect. First, the finite resistance of the transducer fingers increases its significance as the number of finger pairs is reduced. Second, the proportion of bulk acoustic waves to surface acoustic waves generated by an interdigital transducer increases as the number of finger pairs decreases, becoming comparable for few finger pairs [7]. Third, the stray capacitance between the tuning coil and earth is no longer negligible when compared with the transducer capacitance if there are only a few finger pairs. This stray capacitance has the effect of transforming the effective resistance to a lower value. Which of the aforementioned effects is dominant is not yet clear.

Whereas the reflection from a tuned and matched multistrip UDT increases with frequency deviation from resonance, that for a bidirectional transducer decreases. Thus a method of achieving a uniform TTS across the bandwidth is to combine a bidirectional with a unidirectional transducer. Fig. 12 shows the loss and TTS of such a delay line which was iden-

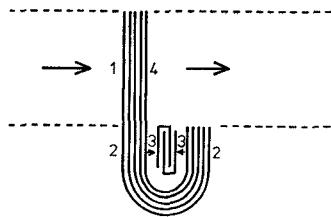


Fig. 13. J-shaped delay line tap.

tical to the 61-MHz unidirectional pair previously described (Fig. 11) except that one U-shaped coupler was absent.

The TTS falls to 20 dB at two points within the passband: it could perhaps be improved further by feeding a narrow-band bidirectional transducer into a UDT of broader bandwidth.

In a practical delay line using bidirectional transducers it is necessary to dispose of the backwards-going wave in order to avoid reflections from the edges of the substrate. We find that no such precautions are needed if UDT's are used.

D. J-Shaped Structures

As a further development of angled couplers, the U-shaped structure can be formed into a J by extending one arm of the U into an adjacent acoustic track. In conjunction with a bidirectional transducer the J forms a unidirectional transducer and, in particular, an efficient tap. Without the transducer it acts as a reflector.

The structure is shown in Fig. 13. Port 3 of the coupler is now split into two opposing halves, the signal leaving each half must reenter the other half. If the coupler has sufficient strips for complete energy transfer, then all of the incident power reemerges at port 1; the structure acts as a reflector. Also, if a transducer is placed centrally in the hook of the J to receive the opposing outputs from port 3, the structure acts as a UDT. This device has the marginal advantage that an offset of $\frac{1}{8}$ wavelength is not required as in the U-shaped UDT, and so the bandwidth is limited only by the intrinsic bandwidth of the coupler. Both of the aforementioned structures are considerably larger than the equivalent U-shaped devices. They have not been experimentally investigated.

A J-shaped transducer with only a fraction of the number of strips required for full transfer may be used as an efficient delay line tap. If the transducers are tuned and matched, then each of a series of such taps will completely absorb a predetermined fraction of the acoustic power traveling along the line. Such a delay line would have minimum loss and negligible acoustic interaction between taps.

V. MULTISTRIP SUBSYSTEMS

The subsystems to be considered in this section make use of the various multistrip components previously described to construct simple acoustic circuits. They do not represent an exhaustive survey of possible combinations but demonstrate some of the advantageous features which can be obtained. These include improved TTS and the ability to transfer a signal from one track to another while simultaneously reversing its propagation direction. This forms the basis of a meander-type delay line. We also show how the capacity of a delay line can be doubled using multistrip components.

A. Reflecting Track Changer

The reflecting track changer and the echo trap described in the following paragraphs both employ the configuration of

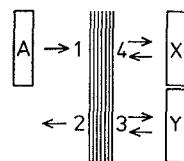


Fig. 14. Acoustic circuit of reflecting track changer and echo trap.

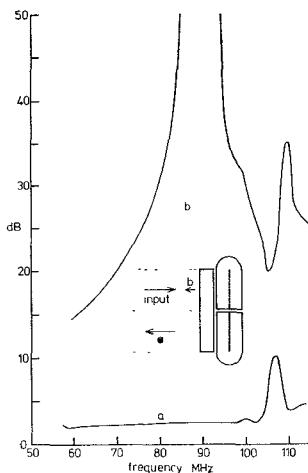


Fig. 15. Reflecting track changer. *a*: Output signal at port 2 relative to input at port 1. *b*: Spurious reflection at port 1.

multistrip components shown in Fig. 14. It consists of a 3-dB coupler and two identical reflecting elements *X* and *Y* both at the same distance from the coupler. A signal incident at port 1 is split by the coupler into two equal signals. *X* and *Y* reflect some fraction of this energy back to the coupler. The signal reentering port 3 maintains the 90° phase advance previously imposed on it, and the coupler recombines the whole of the reflected energy into a single beam that emerges at port 2.

In the reflecting track changer [3], the elements *X* and *Y* are U-shaped multistrip reflectors (Section IV-B). These are introduced with the objective of producing a wide bandwidth, high efficiency, and reflection of the signal incident at port 1 through port 2. By repeated use these reflecting track changers can be used to fold an acoustic beam back and forth on a substrate to produce a long delay line. Since all the paths can be collinear, maximum advantage can be taken of a favorable crystal cut, e.g., the low diffraction loss of YZ LiNbO₃. Further, due to the opposed direction of propagation in adjacent channels, crosstalk can be anticipated to be significantly lower than in helical or spiral guided structures.

The observations reported here were made on a single reflecting track changer. Curve *a* of Fig. 15 gives the decibel ratio between the input signal and the track-changed signal. The 55/107 simple coupler spanned two tracks 4 mm wide. The 60/107 U-shaped reflectors were identical to those described in Section IV-B. Apart from the region of the stopband the loss was between 2 and 3 dB over a bandwidth in excess of 50 percent.

Complete transfer of energy to port 2 can only occur at the frequency for which the simple coupler is exactly $\frac{1}{2}$ transfer length long. At other frequencies some power will return to port 1. Curve *b* shows that the level of this spurious reflection is 30 dB below the input signal over a 22-percent bandwidth centered on 90 MHz. Other spurious signals caused by reflec-

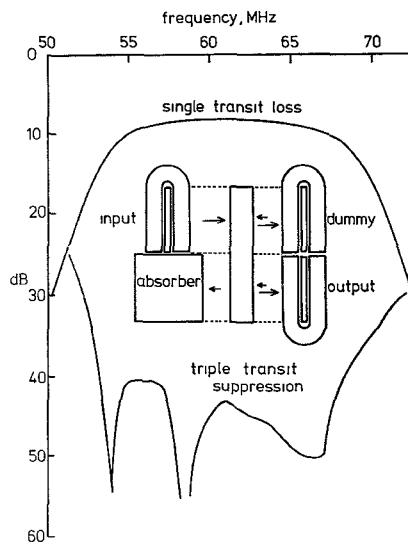


Fig. 16. Echo-trap delay line: STL and TTS versus frequency.

tions from the simple coupler were more than 30 dB down at frequencies below 100 MHz.

The measured insertion loss of the track changer was consistent with the 0.5 dB and 2 dB measured for coupler and reflector, respectively.

B. The Echo Trap

The acoustic circuit shown in Fig. 14 may also be used to remove the triple transit signal from a surface-wave delay line [4].

The input transducer *A* is positioned to feed port 1 of the coupler, and a pair of identical transducers are placed at *X* and *Y*. The output is taken from *Y* and *X* is connected to a dummy load having the same electrical characteristics as the output circuit. The reflections from *X* and *Y* are therefore identical and the coupler diverts the reflected energy to port 2 where it encounters a suitable acoustic absorber. Thus the reflected signal that would normally give rise to the triple transit signal of the delay line is removed at the expense of an extra 3 dB in the STL.

The results given in Fig. 16 are for a delay line incorporating an echo trap and three multistrip UDT's. Each 4 $\frac{1}{2}$ /61/54 transducer was tuned with an identical series coil and the output from transducer *X* was fed into a small variable resistor. 60/86 and 56/86 couplers were used for the U's and central coupler, respectively. It is seen that the triple transit signal is suppressed by more than 40 dB relative to the single transit signal over a 25-percent bandwidth. The number of strips in the central coupler was not critical, and a similar performance could be obtained over a 10-percent variation in this number if the value of the dummy load resistor was adjusted. Small spurious signals were observed that were caused by reflections from the edges of the central coupler, but these were more than 40 dB smaller than the single transit signal.

The results are fully consistent with the previously measured properties of individual couplers and U-shaped reflectors.

Since the reflection from a UDT increases as the frequency deviation from band center increases, the use of these transducers is a very severe test of the echo-trap principle. The device previously described was just capable of a 40-dB TTS over the 25-percent transducer bandwidth. The reflection

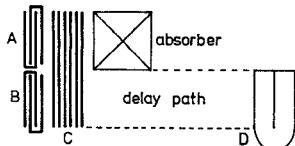


Fig. 17. Use of MSC to double the capacity of delay line. *D* is multistrip reflector.

from a simple bidirectional interdigital transducer decreases as the frequency deviation from band center increases, and TTS greater than 40 dB have been measured over bandwidths limited only by the coupler stopband.

C. Double Length Delay Line

It is possible by using a multistrip circuit to utilize both directions of a delay line sequentially and so double the delay time and hence the available storage capacity of a given substrate. Fig. 17 shows the acoustic circuit employed, *C* is a 3-dB coupler, *D* is a multistrip reflector, and *A* and *B* are transducers. Half of the forward-going signal from the input transducer *A* passes via the coupler to the delay path (this could in fact be a folded delay line employing track-changing elements). The delayed signal returns to the coupler and half of the power is received at the output transducer *B*. If *A* and *B* are at the same distance from *C*, the principle of the echo trap applies and signals reflected from these transducers are diverted by the coupler to the acoustic absorber. Spurious delayed signals therefore have negligible amplitude, but direct RF pickup from *A* to *B* could become significant if the delay line propagation losses are high.

A device closely resembling that used to demonstrate the principle of the echo trap (Fig. 14) was also studied in a double length delay line configuration. The multistrip UDT *X* previously used as a dummy output was used as the input transducer. The output was taken from the adjacent transducer *Y* and the third UDT *A* was left open circuit and so served as an efficient reflector. The transducer had $6\frac{1}{2}$ finger pairs instead of $4\frac{1}{2}$ finger pairs as used in the echo-trap experiment.

The STL had a minimum of $12\frac{1}{2}$ dB and was less than 14 dB over a 16-percent bandwidth. Direct pickup was more than 30 dB below the delayed signal over this bandwidth, and the echo that had traversed the delay path four times was more than 35 dB below the main signal. Reflections were observed that originated at the edges of the central coupler and these were more than 30 dB below the delayed signal.

Even though the delay line may be carrying two oppositely directed signals simultaneously it is still possible to tap one or both of these by making use of the directional property of the MSC (see Sections I-C and IV-D).

VI. DISCUSSION AND CONCLUSION

A. Anticipated Improvements

In the relatively short space of time that work has been in progress on the MSC, we have concentrated on understanding its operation and both demonstrating and exploiting the components which derive from it. There are improvements that can be made to the basic structure that should improve device performance but that we discuss here since they have not been explored in depth.

1) *Increased Coupling Efficiency*: As discussed in [1], K^2 dictates the length of the coupler and in certain circum-

stances can rule out its inclusion in acoustic circuits because of material costs or the transit time of a signal through the device. Recently it has been shown [16] that N_T can be significantly lowered by depositing SiO_2 or ZnO overlayers on the coupler on $YZ\text{LiNbO}_3$. A reduction of the physical size of multistrip components by a factor of 2 is possible. Further, since the reduction in N_T may be interpreted as an increase in K^2 , the optimum bandwidth of transducers also covered by the overlayer is increased [14]. The bandwidth offered by the coupler may then be exploited more effectively.

2) *Reduced Losses*: Surface acoustic wave losses occur in the basic MSC structure due to ohmic losses and mode conversion. Both theoretical and experimental evidence exists for low ohmic losses in 5000-Å-thick Al films on LiNbO_3 due to current flow along the strips. They are significantly less than 0.2 dB in a transfer length [1, sec. V]. Additional losses, of approximately 0.5 dB per transfer length, have been observed even below the stopband.

The reduction of losses in ineffective parts of multistrip components, for example, the bends in U-shaped structures, is more significant because the loss is higher (~ 2 dB per transfer length) and because it occurs in reflectors where low loss is particularly desirable. As stated earlier (Section IV), neither resistive loss nor surface-wave generation is responsible. Bulk mode generation remains a possible source of loss particularly since it may accidentally be favored. A method of reducing this possible loss mechanism is to electrically decouple the metal strips from the piezoelectric substrate by introducing a layer of SiO_2 , say, under the curved strips. The decoupling could be further enhanced by inserting a metal film between the SiO_2 underlayer and the LiNbO_3 surface. Introduction of the SiO_2 underlay has the added advantage that the increase in transfer length caused by the capacity associated with the curved strips is reduced (though this may be offset if a metal film is also included).

Overlays may also reduce loss. Their selective inclusion, i.e., deposition on active parts of multistrip components, can lead to a significant reduction in transfer length and hence a greater ratio of active to inactive parts in, for example, U-shaped structures.

3) *Increased Bandwidth*: Depositing an SiO_2 layer between the coupler and the piezoelectric substrate is expected to increase the bandwidth over which N_T remains essential constant [21]. This will occur since the coupling will fall off with increasing frequency.

An alternative technique is to randomize the strip positions. This could nullify the effect of the stopband. However, as discussed in [1], this technique would present technical difficulties.

4) *Reduced Reflections*: Both in the basic behavior of the coupler and the performance of various multistrip components, it is evident that away from the stopband, reflections from the MSC are small, typically 30 dB or more down. For some applications this level is not acceptable. It can be reduced by "matching" the coupler into the free surface by gradually increasing the coupling. Techniques available include tilting the strips, strip-width weighting [1, sec. IV], and aperture weighting. The latter is simplest and has yielded reductions of reflections of 10 dB.

5) *Track Separation*: In the MSC and the multistrip components embodying parallel tracks that we have discussed so far, the tracks are adjacent. Track separation could be an advantage, particularly in long structures, as a means

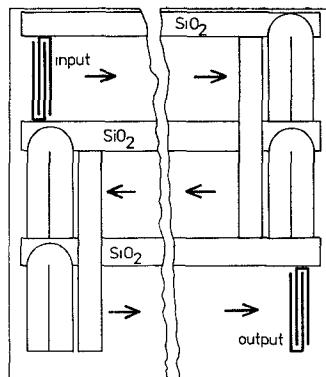


Fig. 18. Meander delay line incorporating reflecting track changers.

of reducing cross talk. One technique of separating tracks is to introduce a nonpiezoelectric underlayer in the portion of the coupler which it is desired to make inactive. With the incorporation of underlays a meander delay line could be constructed, as shown in Fig. 18. The SiO_2 strips serve not only to separate the tracks but to improve the performance of the reflectors as previously discussed.

B. Further Aspects of Multistrip Components

1) *Coupling Between Separate Substrates:* Efficient coupling between two substrates of the same material has been demonstrated (Section II). This raises the question of coupling between dissimilar materials, quartz and lithium niobate, for example, to exploit the high coupling efficiency of one and the zero temperature coefficient of the other. By matching the effective phase velocity (stopbands on each material adjusted to be at the same frequency), a significant amount of power can, in principle, be transferred. However, for the material previously discussed the number of strips necessary might prove prohibitive without recourse to overlays.

2) *Coupling Between Different Waves:* Our discussion has been limited to the coupling between two surface waves. The multistrip coupling principle can be applied to coupling between other identical waves and between dissimilar waves. An essential requirement is the existence of an electric field component parallel to the wave vector if the simple strip coupler is used. The flexibility of the MSC could be utilized to achieve a phase-matched condition through adjustment of the periodic distance in each track. Coupling of surface acoustic waves and space charge waves in a semiconductor is cited as an example of dissimilar waves which could be coupled.

3) *Other Multistrip Structures:* In addition to its role as a true coupler the multistrip structure may also be used as a convenient means of transferring the electric fields associated with a surface acoustic wave from the piezoelectric delay medium to some adjacent structure. This might be a strip of photoconductor, an array of transistor circuits, or any other electrical system which might advantageously interact with the running sound wave but which could not be constructed in the actual path of the beam.

The air gap in the separated medium amplifier [17] may be replaced by a multistrip structure linking piezoelectric and semiconductor to give a robust device of planar construction. Other advantages of this modification are discussed by Tournois *et al.* [19] and Adler [20].

C. Conclusions

In this paper a range of multistrip components have been introduced, the majority of which enjoy exceptional bandwidth. They require neither additional processing nor higher resolution than that used to produce standard interdigital transducers. No external connections are required for their operation. In other words, the multistrip components can be built into the mask used for the photolithography and thereafter require no special treatment. Broken strips or shorts between strips are unimportant compared with the similar problem in transducers; they simply render the individual strips inoperative and can be allowed for in advance by judicious inclusion of additional strips.

The range of components we have described includes reflectors, reflecting track changer, UDT, unidirectional taps, magic T, and beamwidth compressor. One of the most important applications of multistrip components is simply the improvement in delay line performance, and hence a wide range of SAW devices, by reducing the spurious signal level, both bulk and triple transit, and by reducing insertion loss while maintaining the bandwidth. Another important area of application is to long delay lines. Using the reflecting track changer (as in Fig. 18), meander-type delay lines can be constructed and their length doubled (Section V). If necessary, the signal can be transmitted from one substrate to another without significant loss (Section II). In long delay, where length along the propagation track is not at a premium, coupler structures on Bi_2GeO_3 and ceramic become feasible. The virtues of the multistrip taps for forming a tapped delay line include their unidirectionality as well as freedom from spurious signals. Multistrip components could have further repercussions on the design of very long delay lines which require several stages of signal amplification. One may choose to use acoustic wave amplifier structures, but with the availability of broad-band transducers with very low transduction loss it is perhaps simpler to transduce out of the delay line, amplify using transistors, and then transduce again. This approach does not require the development of a new semiconductor technology.

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Acoustic Surface Waveguides—Analysis and Assessment

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Invited Paper

Abstract—The properties of acoustic surface waveguides are reviewed, with particular reference to topographic structures in which guiding is achieved by drastic deformation of the substrate surface. A numerical technique, capable of computing efficiently and with high accuracy the mode spectrum of an anisotropic piezoelectric heterogeneous waveguide of arbitrary cross section, is described. Characteristics of both the ridge guide and the recently discovered wedge waveguide are discussed in some detail.

Techniques for the fabrication of and transduction onto acoustic surface waveguides are discussed, and a preliminary assessment is made of potential linear and nonlinear waveguide applications. A number of experimental devices are described.

I. INTRODUCTION

ACOUSTIC surface waves represent solutions of the wave equation which lead to a concentration of the acoustic energy near to the free surface of a half-space. One can, however, make the stronger statement that the surface *guides* the wave, since it will continue to propagate near the surface even if this should have a small convex curvature [1]. Diffraction for acoustic surface waves is therefore confined to a single plane—that of the free surface. In most devices which have been considered so far one accepts the fact of lateral diffraction as one of the constraints in device design, though one may seek to minimize its effects by the choice of a low diffraction propagation direction

[2]–[4]. There are, however, situations where one would like to obtain guiding action in the plane of the free surface so as to eliminate diffraction completely. A number of types of acoustic surface waveguides have been considered in the past, and detailed accounts will be found in recent review papers [5]–[7]. It is the main purpose of the present paper to review some recent advances in this field, both with regard to the understanding of the characteristics of such guides which has been achieved and to some experiments which bear on possible applications.

Since acoustic surface waveguides are necessarily "open," true guiding action, devoid of leakage, implies that the propagation velocity must, in the guiding region, be reduced to a value less than the Rayleigh velocity and less than the velocity of any coupled bulk waves. Waveguides may be classified in accordance with the means adopted for effecting this velocity reduction. In principle, this can be done by causing a local change in the material itself; for example, by doping in the case of a semiconductor [8], or by depoling in the case of a ferroelectric [9]. There may well be developments leading to applications for such guides. However, the two main categories that have been considered are thin-film guides, in which the velocity reduction is effected by the perturbation effected by the elastic [10]–[12] or electric properties of a deposited thin film [14], and topographic guides, in which the velocity reduction is achieved by a change in the topography of the free surface [7], [13], [15]–[17].

Perhaps the most important single characteristic of an open guide is the strength of the guiding action, as portrayed by the extent to which the acoustic fields extend within the substrate on either side of the guiding structure. It is this which determines the minimum radius of curvature through

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