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contents

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ON THE COVER: The Cobra Judy phased array radar, an Air Force detecting and tracking system for collecting data on foreign strategic ballistic missile tests, is visible at the stern of USNS Observation Island in these three views. Photo courtesy of Raytheon Company.

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SAW Filters for Military and Spacecraft Applications

W. J. Tanski, P. C. Meyer and L. P. Solie
Sperry Research Center
Sudbury, MA

Introduction

Surface Acoustic Wave (SAW) devices are used increasingly in both commercial and military equipment. The small size, rugged construction, wide frequency range and the commercial availability of SAW devices make them an attractive solution to many signal processing problems. In the early 1970's, SAW technology focussed on the development of dispersive delay lines for pulse compression radar systems. Subsequently, as design and processing techniques improved, bandpass filters, resonators and hybrid devices were developed.

Eight Channel Frequency Multiplexer*

There are several techniques that have been used in making SAW multiplexers. The basic problem which all of them have is how to accept a wide input bandwidth and distribute this signal to a number of narrowband output transducers (or vice versa). A straight forward approach of dividing the entire spectrum among N channels would result in a power division loss of 10 log N. However, reduced loss insertion loss can be realized by careful input transducer design and electrical interconnection.[1] An acoustic approach to multiplexing invented and developed at Sperry Research Center is the offset Multistrip Coupler (MSC)[2] multiplexer which transfers a signal between acoustic tracks according to frequency. In this way the power division loss is avoided. The offset MSC is used in the multiplexer to achieve part of the frequency sorting. This technique is flexible and can be used for narrowband or broadband channels. The first device discussed, the 8-channel multiplexer, has 20 MHz wide channels, while in the second device, the 16-channel multiplexer, the channel widths are 1 MHz. The various techniques used to realize the overall frequency response are discussed below.

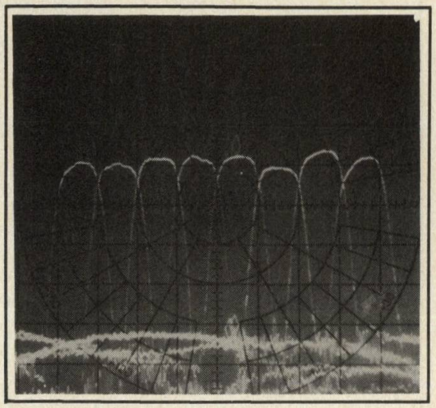


Fig. 1 Frequency response of the 8-channel multiplexer (10 dB/div. vertically); (20 MHz/div. horizontally).

The 50 percent relative bandwidth of the 8-channel device is too large for efficient SAW generation by a single wideband input transducer. Therefore the input signal is fed into a two-way power divider which drives a lower 4-channel multiplexer and an upper 4-channel multiplexer. Each 4-channel multiplexer is on a separate $LiNbO_3$ substrate and in its own flatpack. The problem now is reduced to the design of two 4-channel multiplexers, consisting of four cascaded elements which perform the operations of frequency sorting and filtering:

- an input quadrature phase transducer,
- an offset MSC,
- a pair of fanned MSC's and
- a quadrature phase Flat Exponential Filter (FEF) output transducer.

Each channel has a 3 dB bandwidth of 20.3 MHz and at -45 dB a bandwidth of 38 MHz. The input and output impedance is 50 ohms. The designed insertion loss was -20 dB. Figure 1 is a composite frequency response for all eight channels and shows 28 dB loss. An improved design would reduce the loss to 20 dB and flatten the bandpass response.

16-Channel Frequency Multiplexer

This section describes the use of a SAW multiplexer in a linear frequency-modulated continuous wave (FMCW) radar. These SAW devices are currently used in a low power solid-state millimeter-wave radar for the guidance of antiarmor munitions.[3] Each channel is used as a range cell in the receiver IF section. The rugged construction and small size (less than 0.5 cubic inch for 16 channels) of the SAW devices make them an ideal choice for the missile seeker application.

A linear FMCW radar[4] transmits a periodic waveform in which the frequency varies linearly vs. time. This transmitted waveform is reflected from the target and returned to the radar with a time delay t_d, Figure 2, where $t_d = 2R/C$, where R is the target range and C the velocity of light. This reflected signal is then mixed with the transmitted and gated to obtain the difference signal f_b. Figures 2b and c, are a measure of the range: $f_b = \frac{\Delta f\, t^d}{T}$ and Δf are defined in Figure 4 letting $f_m = \frac{1}{T}$ and rearranging, then $R = \frac{f_b C}{2\Delta f \cdot m}$.

If f_b and Δf are held constant, then the range is inversely pro-

*This work supported by Rome Air Development Center, Air Force System Command under contract F19628-79-C-004.

portional to f_m. A discriminator is used to provide range tracking by controlling f_m. The IF spectrum of f_b is fed into the 16 channel multiplexer to achieve the range resolution of the returns from separate targets. The bandwidth of the IF signal and the characteristics of the narrow-band IF filters are important considerations in determining the radar system range resolution where resolution is the ability of the system to separate the returns from targets at different ranges. This ability will be enhanced if the target return signal at IF has the narrowest possible bandwidth and the bank of narrow-band filters has equivalent performance characteristics. Important considerations for spectral purity of the return signal at IF are transmitter sweep linearity and transmitter FM noise.

Important considerations for the narrow-band filters are center frequency, bandwidth, shape factor, ripple content, and frequency stability.

Typical filter center frequencies are in the range from 30 to 200 MHz. The range resolution is proportional to the ratio of the filter bandwidth to the filter center frequency; therefore, fractional bandwidths on the order of 0.1 to 5 percent are needed. Considering the desired center frequency and the fractional bandwidths required, it becomes very difficult to achieve the performance characteristics with LC-type filters. Crystal filters provide the fractional bandwidths; however, they are limited to the lower center frequencies. Another factor to consider is size or space required for a multiple-channel narrow-band filter bank. The SAW filter results in a major space savings as compared to the LC or crystal filter approach. Other advantages of the SAW filters besides small size, fractional bandwidths, and frequency range are reproducibility, reliability, and relative frequency stability between individual filters in the filter bank.

Figure 3 shows the frequency response of the 16 channel multiplexer. Each channel has a 3 db bandwidth of 1 MHz and a channel separation of 0.5 MHz. The insertion loss is 15 dB.

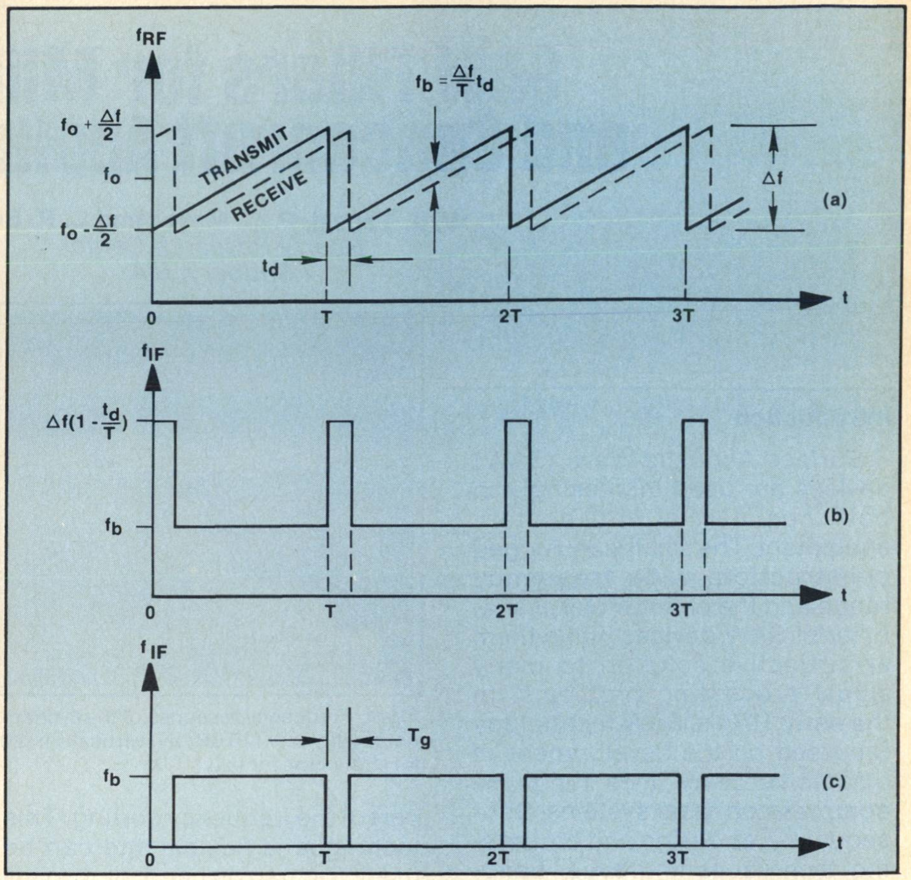


Fig. 2 FMCW radar frequency waveforms. (a) Frequency-time delay for FMCW radar transmitted signal and received signal. (b) Difference frequency in IF circuits without gating. (c) Difference frequency in IF circuits with gating.

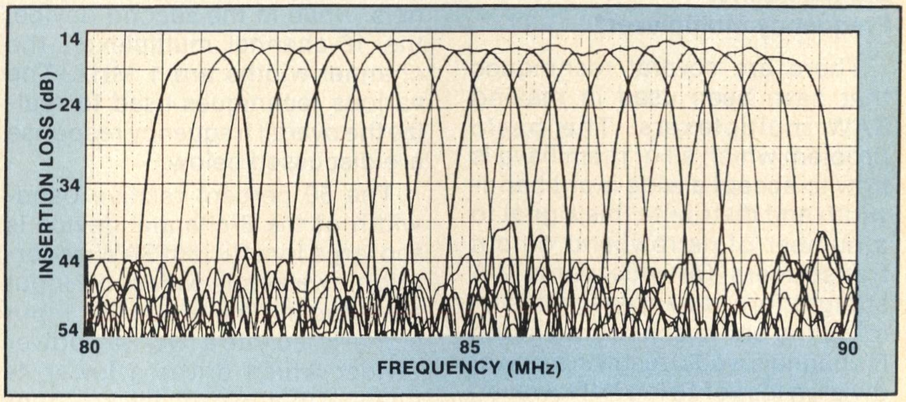


Fig. 3 Output of the 16-channel MSC multiplexer.

Background

The simplest SAW resonator is the surface-wave equivalent of the bulk-acoustic mode crystal resonator which is used extensively in oscillators and narrowband filters. A SAW resonator[5,6] consists of a piezoelectric crystal on the surface of which are placed two distributed reflectors and one or more electro-acoustic interdigital transducers (IDT's). The reflectors are placed to form a cavity that serves to confine the surface wave while the IDT's allow introducing energy into and removing energy from the cavity. The elements of a SAW resonator, configured to form a two-port resonator filter, are shown schematically in Figure 4a. In the two-port device, surface wave energy added to the cavity by the input IDT will travel outward to the reflectors. The reflectors, having reflection coefficients of about 99 percent, return most of the incident acoustic energy to the cavity. A standing wave will form because of constructive interference of the oppositely directed

[Continued on page 56]

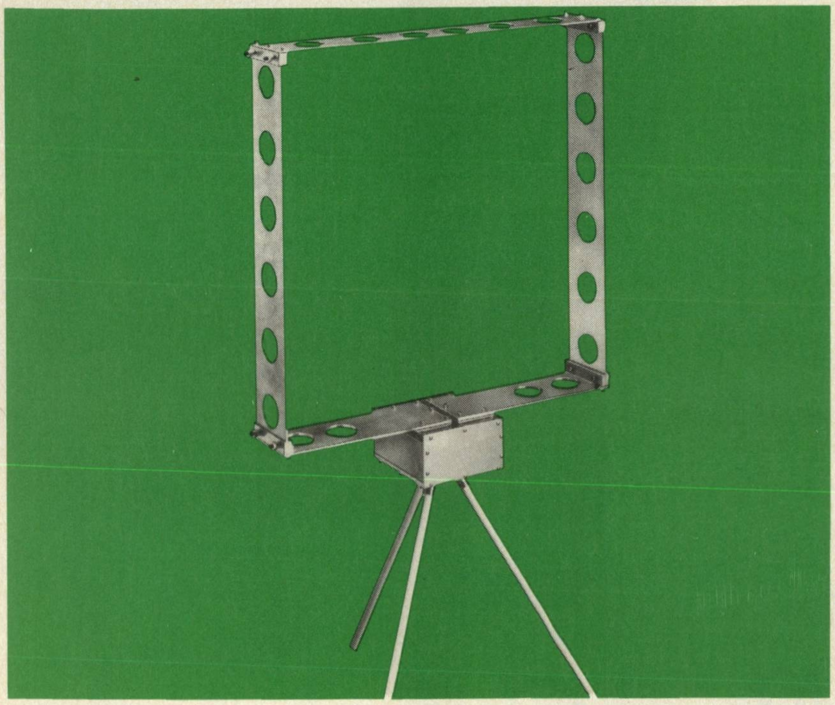
traveling waves when the reflector gratings are an integer number of half-resonant wavelengths apart. This resonant situation is similar to the conventional microwave and optical Fabry-Perot cavities with the important difference that the distributed SAW gratings are efficient reflectors only over a narrow relative bandwidth, on the order of 0.5 percent. Distributed grating reflectors, consisting of a large number ($\simeq$1,000) of very lightly reflecting elements, are necessary in surface wave resonators as these are the only low-loss SAW reflectors known. The output IDT couples to the standing wave and transfers a small portion of the energy in the cavity to the external circuit. In order to suppress unwanted resonances, the IDT's are overlap-weighted and the reflectors may be weighted as shown in Figure 4.

Several physical configurations have been successfully used for the reflectors and transducers of a SAW resonator at frequencies below 400 MHz. However, the recessed-aluminum-transducer/etched-groove-reflector configuration on quartz, shown schematically in Figure 4b, has enabled us to produce devices with Q values approaching the material limit at frequencies as high as 1.43 GHz.[7] The recessed aluminum transducers are used because they have very low acoustic reflectivity on quartz, significantly reducing distortion and losses for SAW resonators[6]. These devices are readily fabricated and physically stable, yielding long-term aging rates as low as 0.1 parts per million (ppm) per year when the device is adequately encapsulated.

SAW resonators are fabricated on the highly polished surface of a quartz crystal; usually the "ST" (45.75° rotated Y) cut for temperature stability. The sequence of fabrication steps are discussed in Reference 6. The groove depth (h) is about 1 percent of a surface-resonant-wavelength (λ_0), that is $h \simeq 0.01\ \lambda_0$. For the ST cut of quartz, with a surface wave velocity $v \simeq 3154$ m/s, the etch depth for a device at 500 MHz, for instance, is about 630 A, with an equal thickness of metal to just fill the grooves. The preferred method of resonator encapsulation is in an evacuated cold-weld-sealed package, Device mounting and encapsulation has a significant effect on device aging and this problem is the subject of much current research.

In order to set a SAW device resonant frequency precisely, we have developed[8] post-fabrication frequency trimming procedure which we find necessary to use to compensate for unavoidable fabrication variances. The physical effect of selectively etching the quartz in a $CF_4 + O_2$ plasma to increase the surface-wave reflectivity of the transducer electrodes (we generally seek to suppress this reflectivity through the use of recessed electrodes) which has the effect of decreasing the cavity resonance frequency. The decrease in frequency is due to the interaction between waves re

[*Continued on page 5*]



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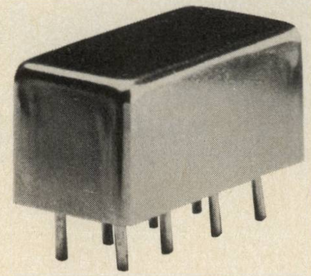

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LO-RF	0.05-500		
IF	0.02-500		
CONVERSION LOSS, dB		TYP.	MAX.
One octave from band edge		6.0	7.5
Total range		7.5	8.5
ISOLATION, dB		TYP.	MIN.
low range	LO-RF	47	40
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[*From page 56*] **SAW FILTERS**

flected from the reflector and those reflected from the transducer. Using this technique we can set a resonance frequency with an accuracy of 5 parts-per-million (ppm) or better and there is no effect on device aging rates.

The equivalent circuit[5] for the two-port/ single-pole filter of Figure 4 is shown in Figure 5 with necessary matching networks. As an example of typical circuit parameters an 840 MHz resonator, with physical parameters as given in Table 1, would have

$R_1 \simeq 675$, $L_1 \simeq 1280\,\mu H$, $C_1 \simeq 7.9 \times 10^{-5}$ pF, $R_o \simeq 15.4$ k, and $C_o \simeq 0.6$ pF.

The required matching circuitry parameters are

$L_s \simeq 2\mu H$, $C_s \simeq 2$ pF and $C_p \simeq 3$ pF.

This resonator design yielded a response shown in Figure 3 R with a loaded Q = 1500 and an insertion loss of 2.4 dB.

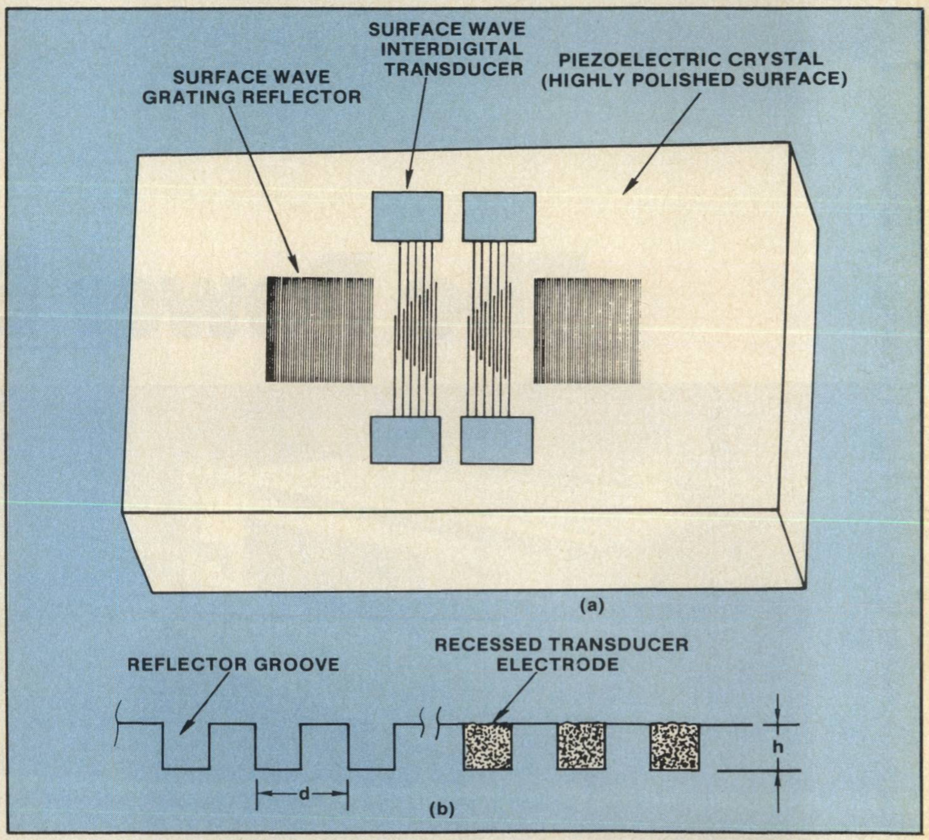

Fig. 4 (a) Diagram of a two-port SAW resonator showing overlap weighted transducers and withdrawal weighted reflectors, and (b) a section of view of the recessed-aluminum-transducer/etched-groove-reflector configuration used to fabricate VHF/UHF and L-band devices.

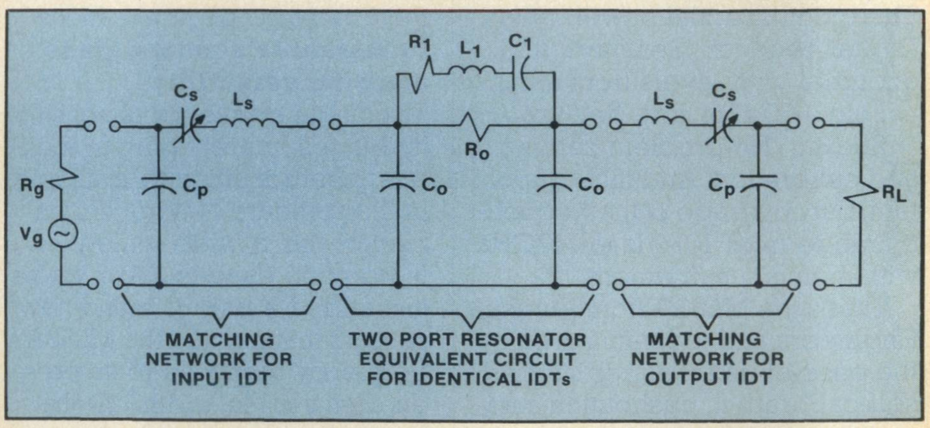

Fig. 5 Two-port resonator equivalent circuit valid in the vicinity of the resonant frequency, and the matching networks.

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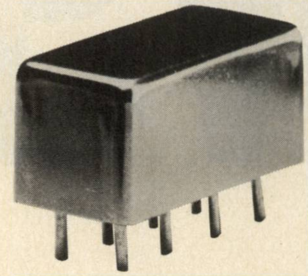
electronic attenuator/ switches



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INPUT 1-200		
CONTROL DC-0.05		
INSERTION LOSS, dB	TYP.	MAX.
one octave from band edge	1.4	2.0
total range	1.6	2.5
ISOLATION, dB	TYP.	MIN.
1-10 MHz IN-OUT	65	50
IN-CON	35	25
10-100 MHz IN-OUT	45	35
IN-CON	25	15
100-200 MHz IN-OUT	35	25
IN-CON	20	10
IMPEDANCE	50 ohms	

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TABLE I

840-MHz Resonator Design Parameters

Wavelength	$\lambda_o = 3.752\ \mu m$
Line Period	$\lambda_o/2$
Line Width	$\lambda_o/4$
Acoustic aperature	$50\ \lambda_o$
Cavity length	$80.0286\ \lambda_o$
Design etch depth	440 A ($h/\lambda_o = .012$)
Reflectors	900 grooves
Transducers	61 Electrodes (30 λ_o)

Multipole SAW resonator filters have been under development[9,10] and this is an area of current research in our laboratory. Energy may be coupled between cavities by several techniques[10], however, the use of coupling-reflectors in conjunction with coupling-transducers is generally preferred since this combination offers the best out-of-band rejection along with suppression of unwanted modes. The frequency response of a six-pole, transducer-coupled 217 MHz filter[11] is shown in Figure 6. The insertion loss of this filter is 6.5 dB with a Butterworth response.

We can fabricate SAW resonators at frequencies up to 1.4 GHz, on quartz, using optical techniques alone. This upper limit is set by the availability of good quality photolithographic masks with linewidths as low as 0.55 micrometers. Using an electron-beam pattern generator to fabricate patterns with 0.25 micron linewidths, the maximum device frequency is about 2.5 GHz. The lower frequency bound on SAW resonators is about 100 MHz due to the large device size at or below this frequency. The maximum possible unloaded Q values are determined by the inherent material losses due to viscosity. The value of surface wave attenuation at 1 GHz for the ST-cut of quartz results in a maximum unloaded Q value given by:

$$Q_u = 10{,}400/F(GHz).$$

Applications

SAW resonators, due to their small size, frequency range of operation, and excellent performance characteristics, may be used to significantly reduce the size, complexity and power consumption of VHF/UHF and L-band subsystems. Specific applications include narrowband filtering,

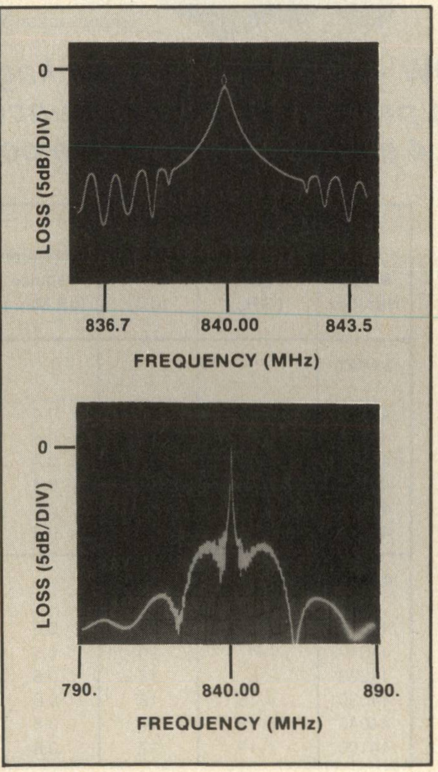

Fig. 6 The frequency response of a two-port 840 MHz resonator used for oscillator stabilization. The loss is 2.4 dB with a Q = 1500. Both wide and narrowband views are shown.

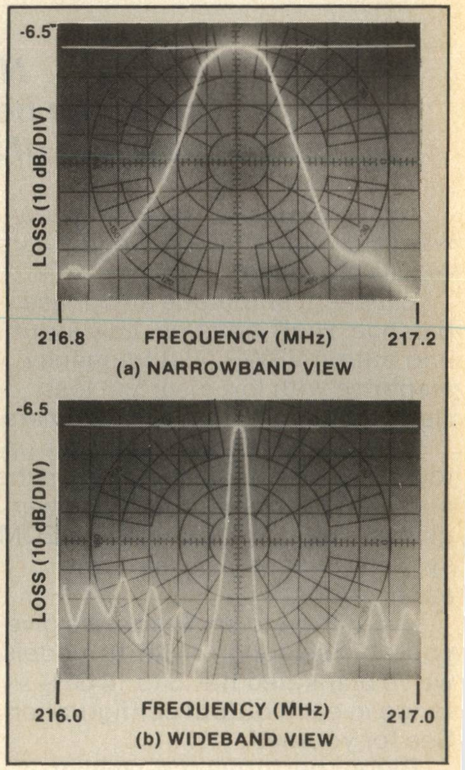

Fig. 7 Frequency response of a six-pole SAW resonator filter. The 3 dB bandwidth is 57 kHz and the 3:60 dB shape factor is 1:3.4 with a minimum rejection of 53 dB.

[*Continued on page 62*]

direct or phase-locked oscillator stabilization. We discuss here two specific applications for which the Sperry Research Center has developed devices thus far.

An 840 MHz two-port SAW resonator filter is being manufactured for use as the oscillator stabilizing element in the MK-92 radar system test set. The device specifications require a center frequency of 840 ± .126 MHz, a linear-monotonic phase shift between the -3 dB points, a maximum 3.5 dB loss, and a minimum Q of 1000 when electrically matched. The devices being produced meet or exceed the required specifications, and the device performance characteristics are listed in Table II. The frequency response of a typical device is given in Figure 7. The SAW oscillator has effected an order of magnitude reduction in volume (19 in^3 for the unit being replaced versus 1.7 in^3 for the SAW oscillator), a deletion of one power supply, and a reduction in power consumption by a factor of 6. Also, this SAW oscillator delivers +26 dBm output power, while

TABLE II

840 MHz SAW Resonator Performance

	Best	Typical
Matched loss	2.4 dB	2.8 dB
Matched Q	2000	1500
Rejection level	18 dB down	16 dB down
Unloaded Q (vacuum)	13,900	12,000
Input power (continuous)	+13 dBm	

utilizing a single RF transistor, and so may be considered a source of 840 MHz power and not simply a signal source. The resonator operates with an input power of +13 dBm, and the single-band noise- to-carrier ratio at 10 KHz offset from the carrier is less than -130 dBc.

A 500 MHz resonator has been designed for use as the noise suppression filter in a low-noise earth-satellite frequency source. The device specifications require a maximum matched insertion loss of 2.5 dB, continuous power handling capacity of +15 dBm and an out-of-band rejection of 22 dB,

TABLE III

500 MHz SAW Resonator Performance

Matched Loss	1.7 dB
Matched Q	2800
Unloaded Q	22000
Rejection	22 dB

minimum. The important device performance features are listed in Table III. These devices must meet stringent environmental specifications, thus they are packaged in a vacuum sealed TO-8 header.

A UHF SAW resonator stabilized oscillator has limitations relating to temperature stability and aging which may be overcome, for very low noise, highly-stable oscillators, by phase locking a SAW oscillator to a multiplied bulk mode crystal oscillator. A SAW oscillator will exhibit higher noise levels than a multiplied crystal source for low Fourier frequencies (F < 5 kHz). However, for higher Fourier frequencies the SAW oscillator operates with lower FM noise levels. Phase-locking allows one to obtain the low close

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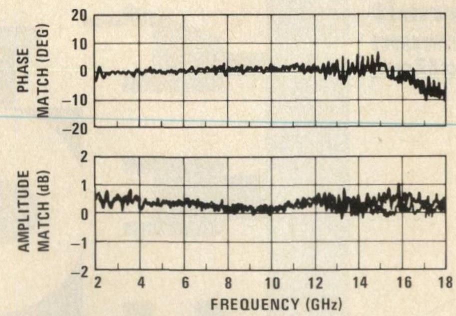

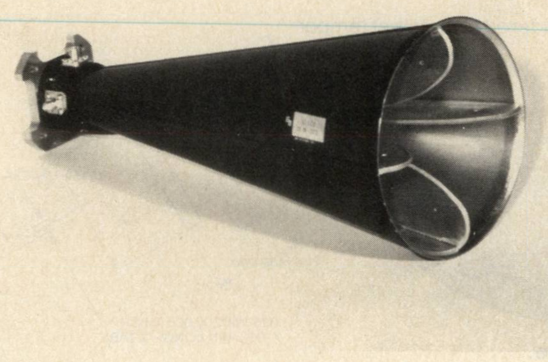

Specifications

Frequency	2 to 18 GHz
Gain	5 to 18 dBi
Polarization	Simul. Horiz. and Vertical
3 dB Beamwidth	60° to 10° nom.
VSWR	2.5:1 max.
Isolation Between Ports	25 dB min.
Phase Tracking Between Ports	±17° max.
Amplitude Tracking Between Ports	±1.3 dB max.
Maximum Power	10 watts
Size	6" Aperture, 13" Long
Weight	4 lbs., 4 oz.

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in noise associated with a bulk crystal oscillator; the low mid-Fourier-frequency-range noise of the SAW oscillator; the long term aging rates and temperature stability characteristics of the bulk oscillator; and relatively high RF output power levels.

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12. Tanski, W.J., R. St. Cyr, P. Dragonetti and E. Kosco, "A Radar System Application of an 840-MHz SAW Resonator Stabilized Oscillator," *IEEE Trans. Sonics and Ultrasonics, SU-28,* May 1981, pp. 146-150. (Also IEEE Trans. Microwave Theory and Techniques, May 1981). ■

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Fast Frequency Hopping Achieved With SAW Synthesizers

Alan J. Budreau, Andrew J. Slobodnik, Jr., and Paul H. Carr
Electromagnetic Sciences Division
Rome Air Development Center
Hanscom AFB

Introduction

A wide variety of both military and commercial systems, including spread spectrum data links, require high spectral purity, low noise, precise, rapid frequency-hopped signal sources, or synthesizers. Hopping speeds of less than a microsecond preclude the use of phase-locked loops. Direct synthesis, in which continuously running sources are selected by a switching matrix and then mixed, yield outputs which can be hopped nearly as fast as the switches themselves. Thus, direct synthesizers can readily change frequency in a fraction of a microsecond. Low-cost, compact, direct synthesizers require filter banks with these same properties. Surface acoustic wave (SAW) filters are a logical choice. A direct synthesizer with a SAW filterbank is shown in Figure 1 which illustrates an approach suitable for generating up to 25 tones. Beyond this, the requirement for a filter per output tone plus the narrow separation between frequencies precludes this simple design. Mixing of the outputs of 2 or more such few-tone synthesizers is a viable alternative and was adopted in all three of the synthesizers to be discussed in this paper. They were all designed to generate approximately 250 tones spaced by 1 MHz in the vicinity of 1.3 GHz. Design goals are provided in Table 1.

Synthesizer results, also given in Table 1, are presented in chronological order. This corresponds to increasing sophistication and miniaturization, as can be seen from the table. Each of these synthesizers logically can be divided into three sections. The source, or comb generator section, is locked to the system clock and produces the ensemble of tones which are then selected and mixed by the remainder of the synthesizer. A property of this synthesizer is that these source tones determine the frequency accuracy of the output which is therefore derived from the clock accuracy. Any fabrication error which shifts filter frequencies from their nominal value results in degradation of the spectral purity of output level, but not in frequency shifts.

RADC Two Mixer Switched SAW Filterbank Synthesizer

By starting with 3 direct synthesizers and mixing together two sets of 9 tones produced as per Figure 1, plus one set of 3 tones generated with phase locked-loop oscillators (PLLO), a capability of 9 x 9 x 3 = 243 tones can be achieved[1] as shown in Figure 2.

— Source Section

The noise and close-in spurious performance, as well as the absolute frequency accuracy of this synthesizer, all depend on the corresponding qualities of the source tones. The stringent spurious and noise requirements of Table 1 put very tough specifications on the source section. Note that two different types of 9-tone comb generators were investigated and used. The comb generator for 360-468 MHz tones, with 12 MHz spacing built by Zeta Laboratories, multiplies up from 1 MHz. It then uses bulk crystal bandpass filters to select the 12 MHz harmonic, and drive a step recovery diode to obtain the desired comb. For the 321-345 MHz tones, with 3 MHz spacing, the mode-locked SAW oscillator[2] (MLSO) shown in Figure 3 was used. A recirculating loop including a SAW delay line, an amplifier to provide net loop gain, and a repressed amplifier to provide nonlinearity was used to directly produce a comb spectrum. While a simple loop consisting of an amplifier and delay line will "lock on" to any frequency which satisfies the requirements of unity gain plus multiple of 2 π phase shift around the loop, the configuration shown in Figure 3 is capable of stable oscillation at multiple frequencies. The directional couplers are used to injection lock from a clock-derived CW reference signal and to couple out the resulting comb signal. The transducers of the delay line are designed to satisfy the phase and amplitude requirements for oscillation at *all 9* required tones. It is the repressed amplifier (an off-the-shelf amplifier operated at greatly reduced bias voltage) that maintains oscillation at all 9 frequencies by its operation in a manner that favors high pulse amplitude signals rather than CW (the Fourier transform of a comb spectrum is a pulse train). Although this SAW based comb generator is smaller, lighter, and consumes less power than the multiply-filter-step recovery diode approach, it is only capable of operation over a few degrees C temperature range at present.

— Switched SAW Filterbank Section

The two high performance 9 channel switchable filterbanks are fabricated on temperature compensated (ST cut) quartz. Each consists of 9 individual SAW filters interconnected to operate as a single-input, single-switch-selected output, device. Each of the filters has a length weighted or apodized (i.e., the overlap of the electrodes is varied) transducer and a withdrawal weighted transducer. The achievement of the out-of-band rejection demonstrated in Figure 5 represents an advance in the state-of-the-art for such filters, and required optimization of many factors and control of second order effects (including bulk and other spurious mode generation, diffraction, and electromagnetic leakage).

Operation as a single-input device required interconnection of the 9 input transducers. The goals of any interconnection or multiplexing technique are:

- insertion loss minimization,
- insertion loss uniformity among the channels, and
- small size and low cost.

Results of our studies[1] indicated that a properly chosen series-parallel interconnection scheme could achieve these goals. In these filterbanks, input multiplexing was achieved by the parallel interconnection of three inductively tuned sets of three series connected transducers. This multiple-inductor scheme results in both low and uniform insertion loss among channels and is especially necessary in the wide percentage bandwidth (26%), high frequency case. Here a mean insertion loss of 25.8 dB with a standard deviation of only 0.75 dB was attained. The strongest spurious tone for the high frequency bank was 57.5 dB below the desired tone, with only about 10% of the spurious tones worse than 63 dB below the carrier. Excellent results were also achieved for the low frequency bank. The average insertion loss of 19.8 dB (including 1.5 dB due to the switch) can be compared to the individual filter loss of 14 dB. A 5 dB improvement over direct power division is achieved. In addition, for the low frequency bank, the worst case spurious was 61.5 dB below the carrier with only about 3% of the spurious tones stronger than 63 dB down.

As shown in Figure 4, the SAW filters are capable of suppressing spurious by 63-65 dB. In addition, under ideal conditions, the switches can achieve 70 dB isolation, particularly for non-adjacent channels. (Note that two switches were investigated: an approximately 2 nanosecond switch[3] having an insertion loss of 16.5 dB, and a 250-400 nanosecond switch from Sanders Associates, Inc. having an insertion loss of 1.5 dB). However, in the finished filterbanks, packaging and adjacent channel switch leakage (possibly also due to packaging) resulted in the levels quoted. A switched filter bank is shown in Figure 5.

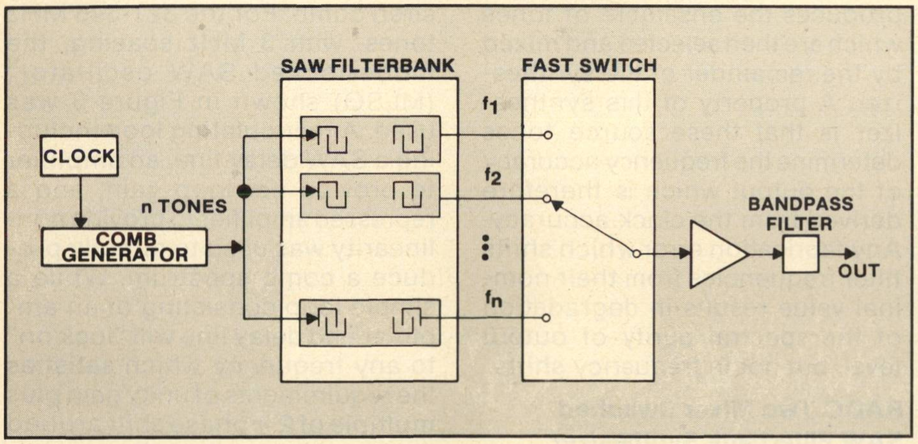

Fig. 1 Schematic illustration of a single channel, n output direct SAW synthesizer.

	GOALS	RESULTS		
		RADC SWITCHED FILTERBANK SYNTHESIZER	HUGHES TWO-MIXER ITERATIVE SYNTHESIZER	TRW FOUR-MIXER ITERATIVE SYNTHESIZER
FREQUENCY RANGE (MHz)	1280-1535	1369-1606	1277-1538	1280-1538
TONE SPACING (MHz)	1	1	1	1
SPURIOUS SIGNAL LEVEL (dBc)	-63	-47	-55	-43
NOISE LEVEL (dBc/Hz)	-120	-110	-120	N/A
SWITCHING SPEED (Nanoseconds)	200	<100	<100	25
SIZE	Small	4x10^{-2} m^3 rack	1.4x10^{-2} m^3 board	2x10^{-3} m^3

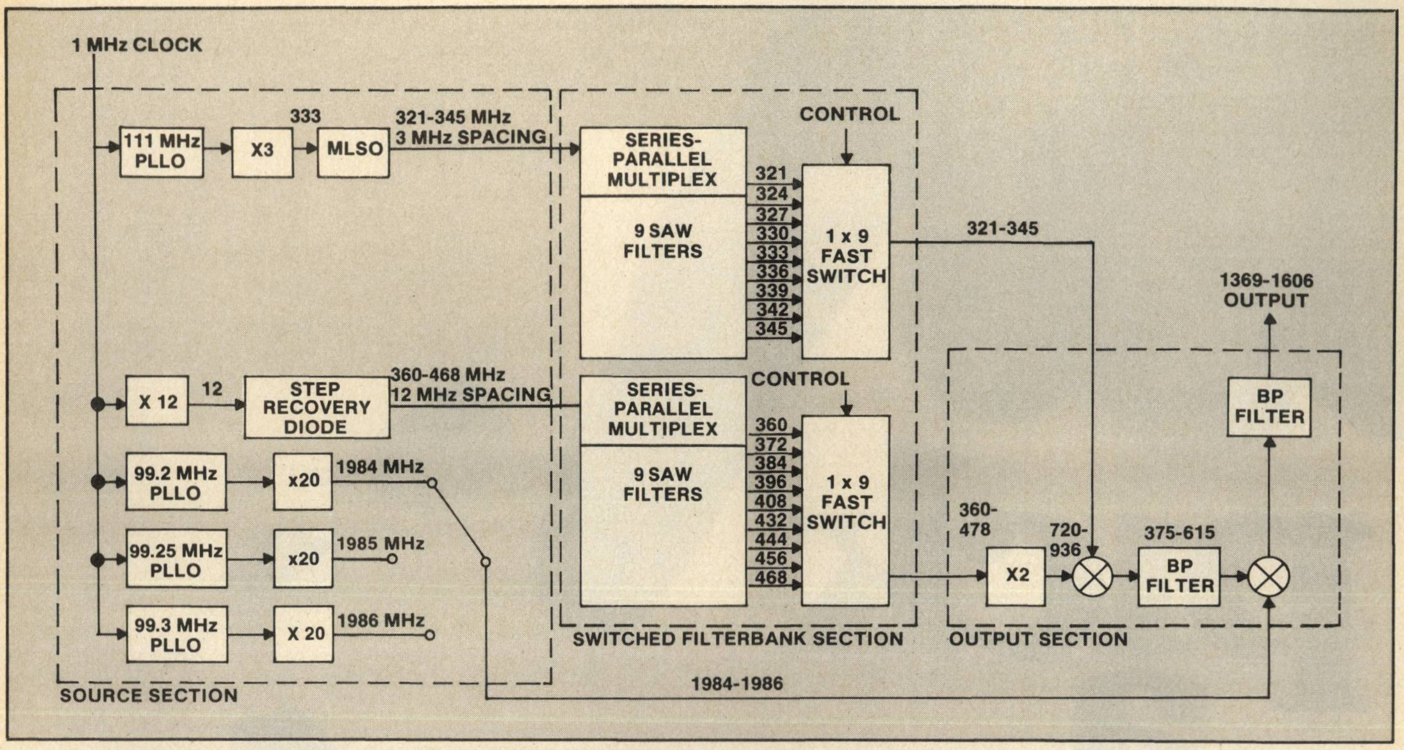

Fig. 2 Block diagram of complete switched filterbank synthesizer.

— Output Section

The output section is the right-hand block of Figure 2. The high performance in spurious rejection and low noise requires that the output section, like the rest of the synthesizer, be carefully designed. Therefore, it is necessary to use high quality double-balanced mixers, filters with high out-of-band rejection, and careful attention to

[*Continued on page 76*]

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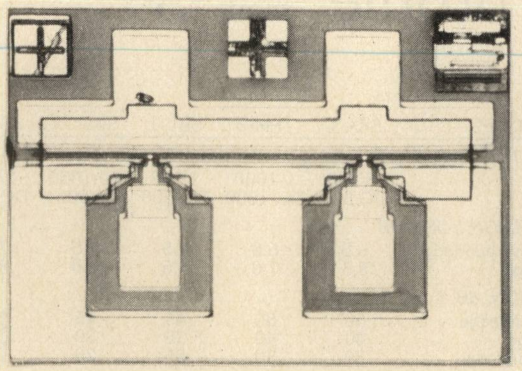

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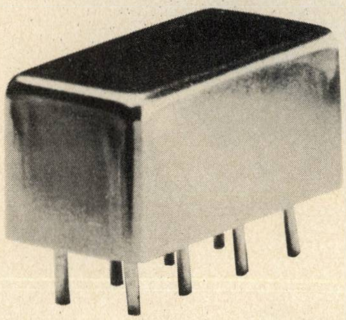

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FREQUENCY (MHz) 0.2-250		
COUPLING, db 19.5		
INSERTION LOSS, dB	TYP.	MAX.
one octave band edge	0.35	0.5
total range	0.35	0.6
DIRECTIVITY, dB	TYP.	MIN.
low range	36	30
mid range	32	25
upper range	25	20
IMPEDANCE	50 ohms	

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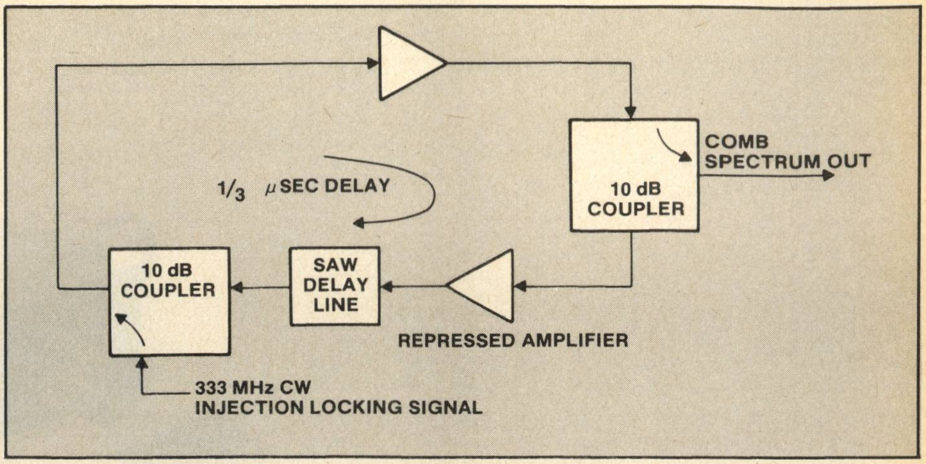

Fig. 3 Block diagram of mode locked SAW oscillator
(Redrawn from figure supplied through courtesy of United Technologies Research Center.)

power levels. Fortunately, these components are not major factors in the synthesizer cost. The output section was fabricated by Communitronics, Ltd.

Whenever two different frequencies are mixed, sum and difference signals plus the input signals, multiples of the input signals, and sums and differences of one input signal with multiples of the other input, are all present. The original choice of the frequencies to be generated and mixed was optimized to minimize the mixer-generated spurious signals.[4] While it is not possible to shift the frequencies so that *all* possible spurious tones are outside the desired band, it is feasible to place one of the input signals so that its higher order undesired products *are* all outside the band of interest, and set this input approximately 20 dB stronger than the other input. The weaker input can then have its higher order spurious fall in-band, but, as they will be extremely weak, they will be adequately rejected by a balanced mixer. This technique was followed.

Low to moderate-priced doublers and mixers were found to have adequate performance provided that power levels were optimized. As in any case where doublers or mixers are used, filtering is required. The doubling of the output of the high frequency switched filterbank avoided the design, fabrication, and leakage problems associated with operation of SAW

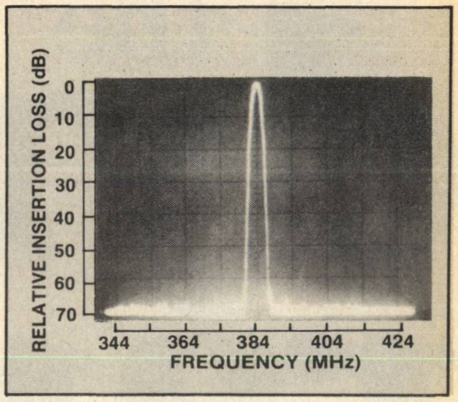

Fig. 4 Frequency response of SAW filter needed to implement the switched filterbank synthesizer.

filters at over 700 MHz. However, as multiplication in frequency by N decreases signal-to-spurious and signal-to-noise ratios by at least 20 log N (i.e. 6 dB for doubling), the requirements for the high frequency switched filterbank were made more difficult by 6 dB.

— Overall Discussion, RADC Synthesizer

The performance of this synthesizer is summarized in the third column of Table 1. Advantages were already listed in the introduction. Although quite good spurious signal suppression was achieved in the filters, a combination of non-ideal doubling, leakage, and the limiting used to level the hopped output degraded the spurious signal performance to approximately 47 dB below the desired signals, as indicated in the table. Figure 6 is the spectrum analyzer trace of an output at

1546 MHz. Although the -63 dBc goal was not realized, the performance actually achieved is sufficient for many synthesizer applications and could be improved by better packaging and shielding.

Hughes Two Mixer Iterative Synthesizer

The iterative approach, in which a relatively small number of tones are repeatedly mixed to yield sum frequencies, then divided by a small integer N (usually equal to the number of tones and most frequently 4) can yield a large number of output frequencies for a small number of source tones. In the Hughes synthesizer,[5] two stages of divide-and-mix are used to obtain 256 output frequencies from the initial 12 tones, as shown in Figure 7. Note that this is the only one of the three synthesizers to use quadrupling, with the inherent 12 dB penalty.

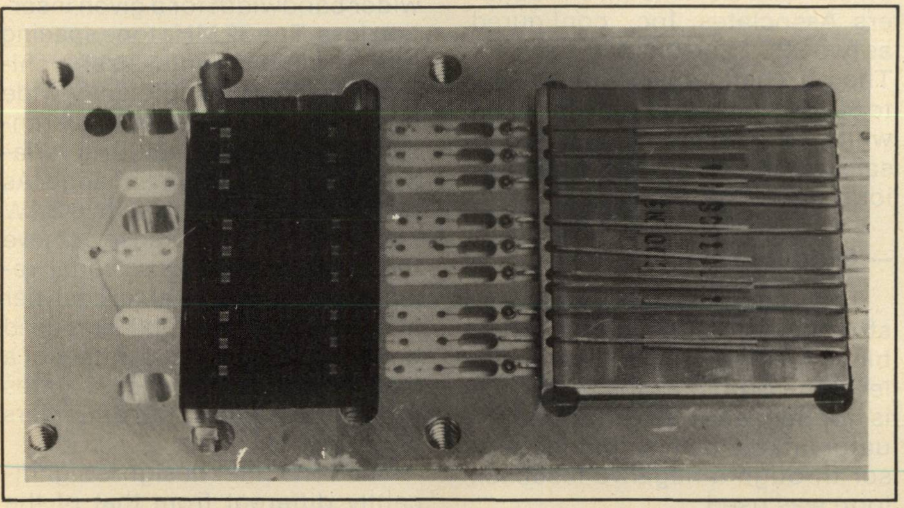

Fig. 5 Photograph of switched SAW filterbank. From left to right: input, tuning inductors (potted), SAW chips with inter-connected input transducers and individual output transducers, output tuning inductors (potted), space for tuning capacitors (not used), test outputs (not used), and switch (2.4 x 2.8 cm).

— Source Section

A step recovery diode comb generator produces a continuum of harmonics from the fundamental up through some roll-off frequency dictated by the circuitry and sharpness (harmonic content) of the generating pulses. Thus, the 240-300 MHz, 4 MHz spaced comb spectrum required for the synthesizer under discussion would have energy at 4 MHz, 8 MHz, etc., up through 300 MHz, which is the 75th harmonic. Clearly, there is great inefficiency in the presence of energy in the fundamental through the 59th harmonic. (The 60th, or 240 MHz, is the first desired). Thus the Hughes synthesizer starts with a CW signal at mid-band (268 MHz) and coherently gates at the desired tone separation (4 MHz) to yield a comb spectrum with its energy centered in the desired band. Another improvement in efficiency is the leveling of the time domain structure of the comb in order to avoid a high peak-to-average power ratio (which increases power handling requirements and therefore amplifier costs). This was accomplished by using a SAW chirped filter placed immediately after the gated source to level the time domain signal while retaining the desired comb structure in the frequency domain.

— SAW Filterbank

As in the RADC synthesizer, Hughes used ST cut quartz filters

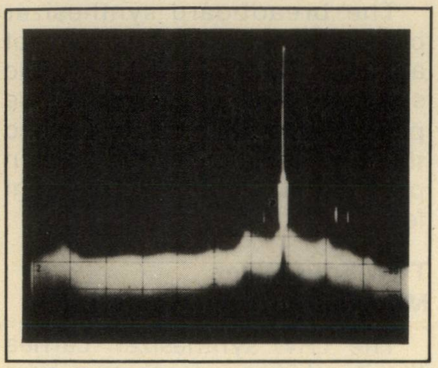

Fig. 6 1546 MHz output tone from switched filterbank synthesizer. 10 dB/div vertical scale. 30 MHz/div horizontal scale.

[Continued on page 78]

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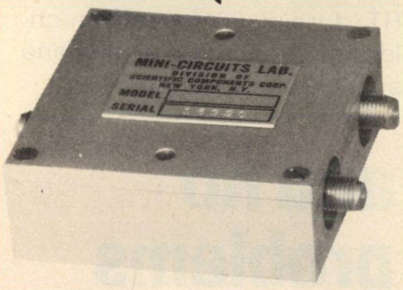

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LO, RF	1.5-4.2		
IF	DC-0.5		
CONVERSION LOSS, dB		TYP.	MAX.
Total range		7.0	8.5
ISOLATION, dB		TYP.	MIN.
1.5-2.0 GHz	LO-RF	25	20
	LO-IF	18	10
2.0-3.7 GHz	LO-RF	25	17
	LO-IF	18	10
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to achieve low temperature sensitivity. However, rather than pursue very high out-of-band rejection in a single filter, this filterbank used pairs of SAW filters cascaded with a simple matching network. Each filter consisted of a length weighted transducer and a withdrawal weighted transducer. Interconnection of the 12 filter channels was accomplished through a constant-k ladder,[6] with the advantage of allowing each input to have a grounded electrode, thereby reducing capacitive electromagnetic feedthrough. In addition, the constant-k ladder operates well in the presence of the filter parasitic resistance.

— Switch

As shown in Figure 7, the switch must partition the twelve CW signals from the filterbank to the three mixer ports in two subsets of 8 and one of 4. A multipole RF switch was obtained from Sanders Associates, Inc., configured as two 8PST and one 4PST switch. The off channel isolation was 60 to 80 dB, compared with the 90 dB which would have been necessary to achieve the full 63 dB spurious signal rejection.

— Output Section

The output section was constructed of commercially available amplifiers and doublers. To level the output as the frequency is hopped, a variable diode attenuator linked to an amplitude sensor through a negative feedback loop was used.

— Overall Discussion, Hughes Synthesizer

The breadboard synthesizer built to demonstrate this concept achieved approximately -55 dBc spurious levels, and otherwise performed well. It was estimated that the cost of this synthesizer, in production, would be 50 percent less than the cost of the switch.

TRW Four Mixer Modular Iterative Synthesizer

The TRW synthesizer[7] carries the iterative approach to a goal of: yielding 256 tones with only 4 source frequencies. By judicrous design of the source section, only two source oscillators are required. Further, the use of 4 identical modules as indicated in Figure 10 made it feasible to develop custom RF LSI chips, leading to the smallest and possibly lowest cost of the three synthesizers.

— Source Section

Phase-locked loop oscillators at 512 and 544 MHz are coupled to the 40 MHz clock. These two tones are combined and amplified nonlinearly to generate the additional required 480 and 576 MHz tones as intermodulation products to yield the 4 tone comb spectrum.

— SAW Filterbank

Unlike the previous two synthesizers, which used ST cut quartz to achieve low temperature dependence of the filter frequencies, the TRW unit uses YZ lithium niobate, which has higher piezoelectric coupling. This allows wider bandwidth for a given insertion loss. The 32 MHz tone spacing inherent in this synthesizer architecture allows a sufficiently wide passband to incorporate tolerances for both manufacturing variations and temperature change. As with the Hughes design two SAW filters in series are used to achieve the required spurious signal rejection at a reasonable cost per filter; however, an amplifier is used between the individual filters of a pair. Since this approach requires four pairs of filters, four amplifiers are required.

The TRW filter design is significantly different from that of the previous synthesizers. Here each filter has an input transducer of thinned single electrodes and an output transducer of length weighted double electrodes operated at their third overtone. The length weighting has a sin x/x function which yields a rectangular bandpass. Distribution of the input comb to the four channels uses a series-parallel interconnection similar to that of the RADC synthesizer.

— Iterated Module

This module accomplishes the functions shown in the dashed block of Figure 8. Four input frequencies and another from the

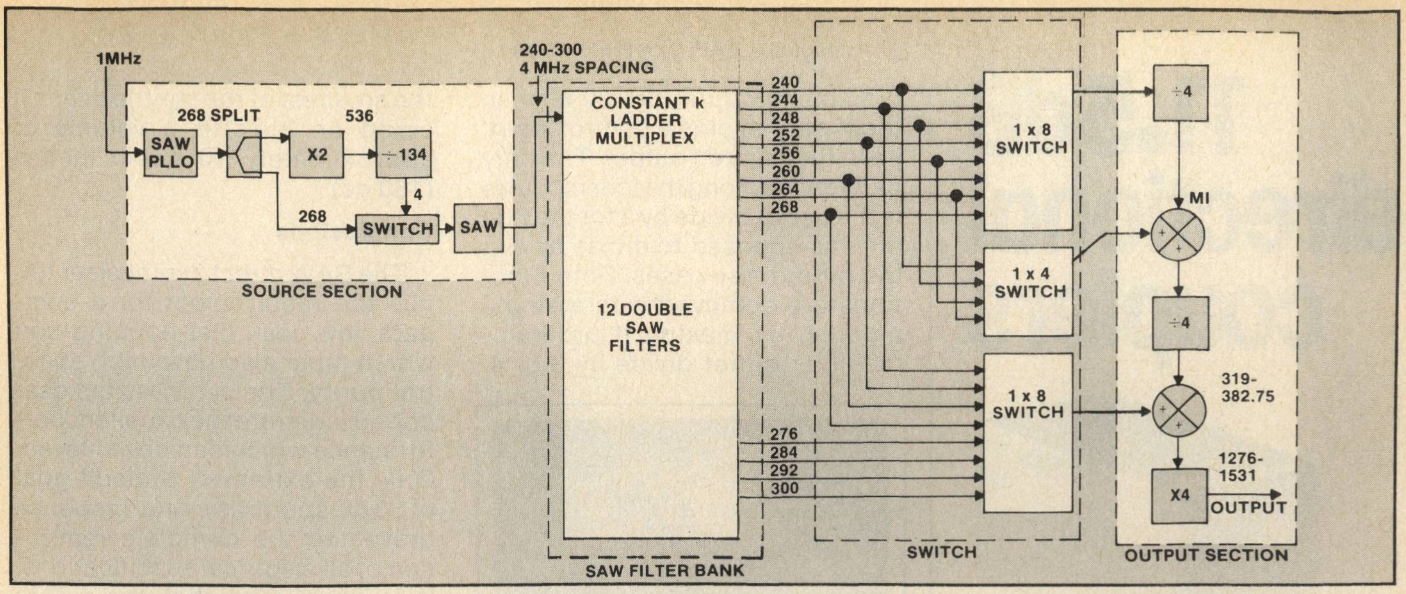

Fig. 7 Block diagram of Hughes two-mixer iterative synthesizer. (Redrawn from figure courtesy of Hughes Aircraft Co.)

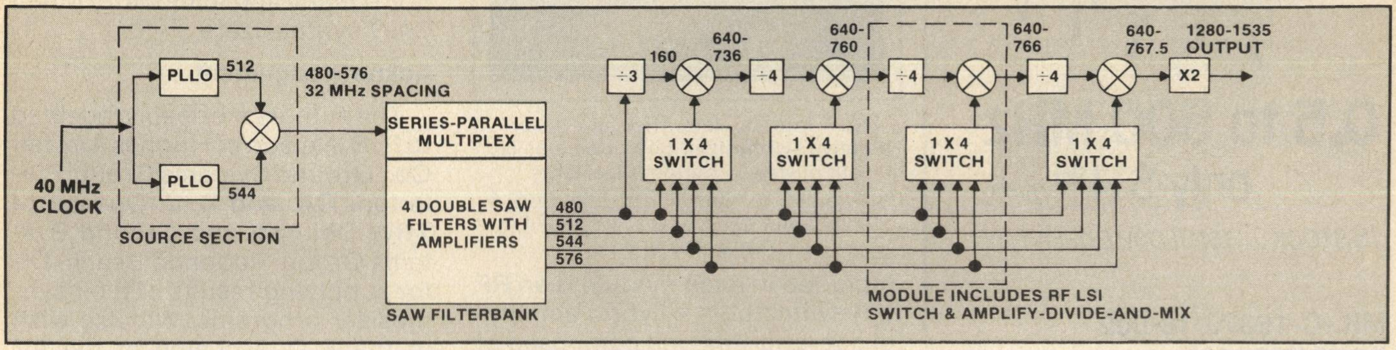

Fig. 8 Block diagram of TRW four-mixer iterative synthesizer. (Redrawn from figure supplied through the courtesy of TRW).

[Continued on page 80]

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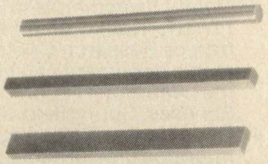

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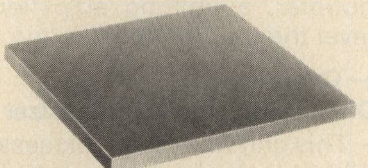

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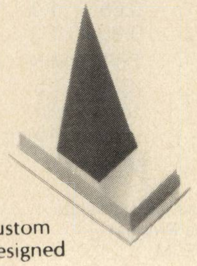

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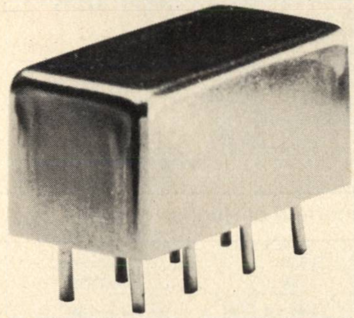

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FREQUENCY (MHz) 0.5-500		
COUPLING, dB 11.5		
INSERTION LOSS, dB	TYP.	MAX.
one octave band edge	0.65	1.0
total range	0.85	1.3
DIRECTIVITY, dB	TYP.	MIN.
low range	32	25
mid range	32	25
upper range	22	15
IMPEDANCE	50 ohms.	

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previous module (or 480 MHz in the first one) plus a control input, yield the desired output. The only difference among the four modules is the use of divide by 3 for the first one, as opposed to divide by 4 in the other three cases. Both functions are achieved with a single module, by making it programmable to either divide by 3 or 4.

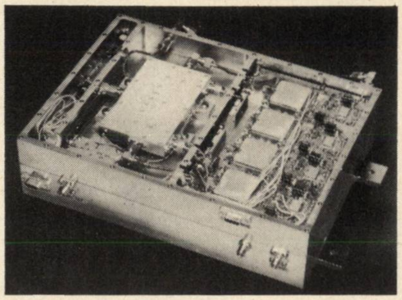

Fig. 9 Interior view, TRW four-mixer iterative synthesizer. Source section is out of view in bottom slice; large rectangular package is SAW filterbank; 4 square packages on right are amplify-divide-mix unit.

Included in each module is an RF LSI switch plus a hybrid amplify-divide-and-mix unit composed of three RF LSI chips. The 1 x 4 switch consists of an amplifier in each of the four channels, fabricated in differential circuitry. It has less than 25 nanoseconds switching time. In their present versions neither the switch nor the amplifier-divide-and-mix hybrid have sufficient isolation to achieve the overall -63 dB synthesizer spurious signal rejection. Both could be improved with further development.

— Output Section

The final section consists of a doubler, a limiting amplifier to level the output, and filtering.

— Overall Discussion, TRW Synthesizer

This unit was highly successful, except for the higher than desired spurious levels. As stated, improvement is feasible by changes in both the layout on each chip and the packaging. The outstanding feature of this approach is the use of four identical modules thereby achieving a factor of four economy of scale for quantity production, and making the use of custom chips practical. For this reason, the TRW unit is inherently the smallest of the synthesizers; it could be built in a volume of approximately 10 cubic inches (160 cc).

Conclusions

The SAW direct synthesizer fulfills the requirement for a compact, low cost, fast-hopping-unit which must also have high spectral purity. The synthesizers described illustrate the excellent performance which can be achieved. Only the extremely difficult goal of 63 dB spurious signal response prevented the complete realization of all electrical specifications. It is concluded that the future holds many applications for direct SAW synthesizers.

Acknowledgement

The authors are deeply indebted to R. B. Nelson of Hughes Aircraft Co., Ground Systems Group, Fullerton, CA, and D. J. Dodson of TRW Defense and Ground Systems Group, Redondo Beach, CA for supplying results of their synthesizer programs. We also wish to thank D. McLaine of the Air Force Materials Laboratory and H. Schecter of the Electronic Systems Division for their longtime encouragement and support.

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MW Signal Processing with Magnetostatic Waves and Modes

A tutorial overview of magnetostatic volume and surface wave propagation in ferrite crystals.

F. R. Morgenthaler
Dept. Of Electrical Engineering
and Computer Science
MIT, Cambridge, MA

Introduction

One of the motivations for exploring the use of magnetostatic waves (MSW) in microwave signal processing devices arises from the fact that MSW operation at RF frequencies is possible in the range of 1-10 GHz and beyond. This is in marked contrast to the situation with surface acoustic wave (SAW) devices that usually operate on the IF signal.

The basic difference between the frequency dispersion of SAW and MSW is depicted in Figure 1 which illustrates the reduction of the elastic wavelength for increasing frequency. Because the Rayleigh wave velocity, V is on the order of 3×10^5 cm/sec., the wavelengths shrink to submicron size for S band frequencies and above. This makes fabrication of interdigital transducers with half wavelength spacing progressively more difficult; moreover, since the wave penetration into the substrate is on the order of λ, imperfections can cause catastrophic deterioration of SAW device performance — even beyond the increased insertion loss expected due to bulk attenuation.

Although the situation is alleviated somewhat by employing a "fast" substrate for which V (and therefore λ) is larger, other material properties may have to be sacrified and, in any event, piezoelectric materials providing very large increase in V are not available.

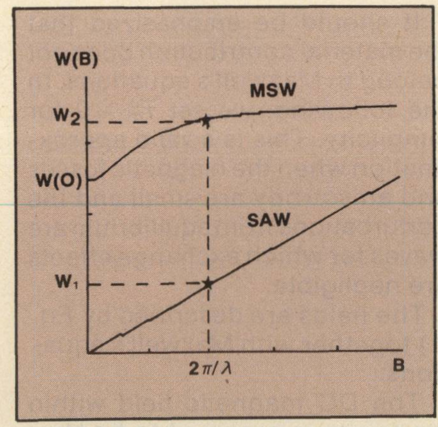

Fig. 1 Comparison between SAW and MSW normalized frequency dispersion. For a given wavelength λ, the MSW frequency ω_2, can be much higher than the SAW frequency w_1. Moreover, since $\omega(o)$ depends upon the dc bias, ω_2 is magnetically tunable.

In contrast, frequency dispersion of MSW modes of a thin ferrite film can have the form shown, with the intercept $\omega(0)>0$ tunable by means of a DC magnetic bias. This implies that λ can be adjusted to any convenient value so long as it is small enough to provide magnetostatic slow wave character. The result is that the same MSW transducer can be magnetically tuned to operate at a variety of frequencies in the GHz range.

It should be emphasized that MSW and SAW are complementary rather than competitive technologies. For the former, time delays are typically in the tens to hundreds of nanoseconds but are achievable over gigahertz bandwidths.

Here we present an overview of magnetostatic volume and surface wave propagation in ferrite crystals, especially thin rectangular films placed between conducting ground planes. When the uniform bias is parallel to one of the three principal axes of the film, the corresponding MSW is either a surface or volume wave; the former is termed MSSW, the latter MSFVW or MSBVW according to whether it is a forward or backward volume wave.

The connection between MSSW propagation in a rectangular film and magnetic resonance in a small spheroid is emphasized in order to clarify the physics of both and because it provides a key to understanding the manner in which nonuniform bias fields can alter and hence control the frequency dispersion of MSW.

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The prospects for controlling the frequency, speed and energy distribution of magnetostatic modes by means of nonuniform bias fields are discussed; especially the prospects for nondispersive MSW delay. Experiments are also reviewed that demonstrate the guiding of MSSW around bends by the use of an in-plane bias with nonuniform orientation.

Model for a Magnetized Ferrite

Magnetic materials with very low magnetic resonance and dielectric losses allow microwave energy to penetrate and strongly interact with the precessing magnetization vector. Such precession about a DC magnetic bias field is similar to that of a child's top when it is spinning under the influence of gravity, except that the precession occurs at microwave frequencies.

A number of practical materials have such properties. Commonly called ferrites, they are oxides of the ferromagnetic metals which may also contain ions of one or more nonmagnetic atoms. The most important of these is yttrium iron garnet, $Y_3Fe_5O_{12}$ (commonly abbreviated YIG). In recent years, high quality single crystal films, grown by liquid phase epitaxy (LPE), have become available. The improved quality of these films, (necessary for MSW applications,) is owed in part to the development of LPE grown garnet films for magnetic bubble memories.

The bound electrons of a magnetic atom give rise to angular momentum vector $\overline{J}$ made up of both spin and orbital contributions. Associated with each $\overline{J}$ there is a magnetic moment vector $\overline{\mu} = \gamma\overline{J}$ where γ is the gyromagnetic ratio given by 1/2 the product of the electron charge to mass ratio and the Lande g-factor. In most practical ferrites, the orbital contributions cancel and g = 2.

Because the individual magnetic moments are strongly coupled to their nearest neighbors, a spontaneous magnetization vector $\overline{M}$ exists for temperatures below the Curie temperature; there must also be a macroscopic angular momentum density given by $\overline{J} = \overline{M}/\gamma$.

In the absence of an applied magnetic field, a ferrite specimen will subdivide into many small domains in which the direction of the moments alternates. The transition regions between oppositely magnetized domains can oscillate at microwave frequencies thereby absorbing power. Such losses can be prevented by applying a DC magnetic field strong enough to wipe out the domain pattern and we assume this has been done. In the magnetically saturated sample, the magnitude of $\overline{M}$ (and hence also $\overline{J}$) is a constant at every point within the material.

The basic equation of motion governing the magnetization in a saturated ferrite is Newton's LAW $d\overline{J}/d\tau = \overline{\tau}$

where $\overline{\tau}$ is the total torque density composed of the electromagnetic part $\overline{\tau}_{e\mu} = \mu_o \overline{M}x\overline{H}$ and a material part $\overline{\tau}_{\mu}$ arising from interactions between the magnetic moments (exchange) as well as those involving the lattice (anisotropy, and magnetic loss). Since the spacing between moments is fixed and only the spin direction can change, $\overline{\tau}_m$ can be written as μo $\overline{M}x\overline{H}^m$ and with $\overline{H}^E = \overline{H}+\overline{H}^m$, see A -Eq. (1).

It should be emphasized that the material contribution does *not* belong in Maxwell's equations. In the following, we set $\overline{\tau}_m = 0$ for simplicity. This is a valid approximation when the magnetic losses and anisotropy are small and the perturbations from equilibrium are waves for which exchange effects are negligible.

The fields are described by Eq. (1) together with Maxwell's equations.

The DC magnetic field within the ferrite is assumed to be $\overline{H}_o = i_z H_z$; the magnetization vector in static equilibrium $\overline{M}_O = i_z M$.

When deviations from static equilibrium occur, there exist in addition, small signal fields $\overline{m}$, $\overline{h}$, $\overline{e}$ where $\overline{m} \cdot i_z = 0$.

Substitution into Eq. (1) and the neglect of $\overline{m}$ x $\overline{h}$ yields, see A- Eq. (2)

where $\omega_Z = -\gamma \mu_O H_Z > 0$ and $\omega_M = -\gamma \mu_O M > 0$.

Notice that if $\overline{h} = 0$ (no electromagnetic coupling), Eq. (2) has the simple solution, see A- Eq. (3) where ϕ can be an *arbitrary* function of position. Because of the characteristic precession frequency, one might expect a resonant response from the material whenever an applied RF excitation occurs at the frequency ω_z which corresponds to a frequency of 2.8 MH_Z/0e. The actual situation is more complicated because an applied $\overline{h}$ field in general causes the material to produce a reaction field that shifts the resonance frequency.

Equation 2 implies that circularly polarized components of $\overline{m}$ and $\overline{h}$ are linearly related by scalar magnetic susceptibilities $m^{\pm} = \chi^{\pm} h^{\pm}$ where $m^{\pm} = 1/2(m_x \pm jm_y)$, $h^{\pm} = 1/2(h_x \pm jh_y)$, $\chi^{\pm} = 1/Z \pm \Omega$, and we have defined dimensionless quantities $Z = \omega_z/\omega_M$ and $\Omega = \omega/\omega_M$.

It is often convenient to utilize two alternate susceptibilities, χ and κ, defined by $\chi = 1/2(\chi^+ + \chi^-)$ and $\kappa = 1/2(\chi^+ - \chi^-)$.

For general plane waves of frequency ω, the complex $\overline{e}$, $\overline{h}$ and $\overline{m}$ all are proportional to $\exp(-\overline{\gamma}\bullet\overline{r})$ where $\overline{\gamma} = \overline{\alpha} + j\overline{\beta}$. When $\overline{\alpha} = 0$, the wave is uniform.

From Maxwell's Equation it follows that, see A- Eq. (4).

Whenever either $|\overline{y}\bullet\overline{y}| >> w^2 \varepsilon\mu_o$, where ε is the dielectric permittivity of the ferrite,

$\nabla x\overline{h} = \omega^2 \varepsilon\mu_O(\overline{\gamma}x\overline{m})/(\overline{\gamma}\bullet\overline{\gamma}+\omega^2 \varepsilon\mu_O)$ is small. The magnetostatic approximation (MSA), defined by setting ∇ x$h = 0$, is then valid and $\overline{h} \simeq -\nabla\psi$ where ψ is the RF magnetostatic potential. Although the electric field energy contribution $(\frac{1}{2}\varepsilon|\overline{e}|^2)$ to the wave energy density is negligible, it should be recognized that the wave power flux remains given by $\overline{e}x\overline{h}$ and is in general nonzero.

For a uniform dc bias field along z, the magnetostatic potential ψ satisfies Walker's Equation[2], as shown in A- Eq. (5)

In the air regions immediately outside the ferrite, $\chi = 0$ and Eq. (5) reduces to Laplace's Equation.

From Eq. (5) the normalized frequency of a single magnetostatic wave is, see A- Eq. (6)

For a uniform plane wave, $\overline{\gamma} = j\overline{\beta}$ and Eq. (6) can be expressed as

$\Omega^2 = Z^2 + Z\sin^2\theta$ where θ is the angle between $\bar{\beta}$ and the z-axis. The normalized frequency is strongly dependent upon the direction of propagation but not upon the magnitude of $\bar{\beta}$. We next consider the effect of transverse boundary conditions on MSW propagation[3].

— Magnetostatic Forward Volume Waves (MSFVW)

For the Figure 2 geometries (with $x_1 = x$, $x_2 = y$, $x_3 = z$) and neglecting width variations

$\gamma_x = 0$, $\gamma_y = j\beta$, $\gamma_z = \pm jk_z$

Therefore from Eq. (6), A- Eq. (7).

A pair of such propagating waves with opposite values of k_z creates a standing wave along z that satisfies the boundary condition at the film surfaces $z = \pm d/2$, provided in A- Eq. (8).

The potential is either an even or odd function of z. When D = 0, $k_z = p\pi/d$ with p = 0, 1, 2,. . .

These modes (shown plotted in Figure 3a for the case $D = \infty$, Z = 1) are termed forward volume waves because they all have positive group velocities, trigonometric thickness variations and lie within the volume wave manifold

$Z \leq \Omega \leq \sqrt{Z(Z+1)}$

For the lowest order even mode without ground planes, the dispersion for $\beta d << 1$ is approximately $\Omega \simeq 1/4\beta d$. Therefore the group velocity is given by $v_g = \omega_M \partial\Omega/\partial B \simeq 1/4\omega M\, d$ and is controlled by the product of saturation magnetization and film thickness. (For a 10 μm YIG film, $v_g \simeq 8 \times 10^6$ cm/sec.

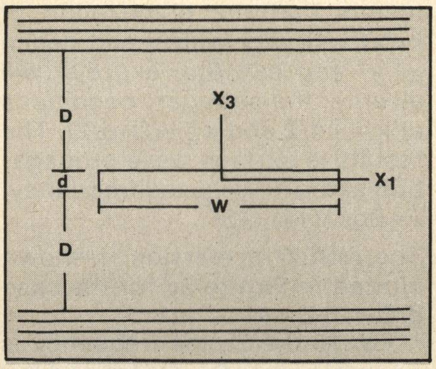


Fig. 2 Transverse section of a thin ferrite strip placed between ground planes. Longitudinal propagation of MSFVW, MSBVW and MSSW results when respectively, the dc bias is parallel to the x_3, x_2 and x_1 axis.

— Magnetostatic Backward Volume Waves (MSBVW)

For the Figure 2 geometry (with $x_1 = x$, $x_2 = -z$, $x_3 = y$), again neglecting width variations

$\gamma_z = 0$, $\gamma_y = \pm jk_y$, $\gamma_z = j\beta$

Therefore, from Eq. (6) A- Eq. (9)

A pair of waves propagating with opposite values of k_y creates a standing wave along y that satisfies the boundary conditions at the film surfaces, $y = \pm d/2$, provided A- Eq. (10).

The potential is either an even or odd function of z. When D = 0, $k_z = p\pi/d$ with p = 0, 1, 2,. . .

The modes shown plotted in Figure 3b for the case $D = \infty$, Z = 1, are termed backward volume waves because they have negative group velocities, trigonometric thickness variations and lie within the volume wave manifold.

— Magnetostatic Surface Waves (MSSW)

For the Figure 2 geometry (with $x_1 = z$, $x_2 = y$, $x_3 = -x$) and still neglecting width variations, $\gamma_z = 0$ and $\gamma_y = j\beta$.

From Eq. (6) it appears that $\Omega^2 = Z(Z+1)$ independent of γ_x and β. For volume waves this is true because $\gamma_x = jk_x$ and $-\gamma_x^2 + \beta^2 > 0$. However, for nonuniform plane waves satisfying $\gamma_x = \beta$, Eq. (6) is indeterminate, and we must proceed differently. To clarify the physics involved, let us first consider the geometry that consists of a single air-ferrite interface at x = 0 that bounds a ferrite half-space (x>0). The magnetic field is taken parallel to the z-axis.

Because $|\gamma_x| = |\beta|$ in both regions, we take, see A- Eq. (11)

In the ferrite region $h_f^{\pm} = (-|\beta| - \beta)\ 1/2\ \psi_f$ whereas in the air, $h_a^{\pm} = (|\beta| - \beta)\ 1/2\psi$. The boundary conditions require continuity of ψ and $(m_x + h_x) = (1+\chi^+)h^+ + (1+\chi^-)h^-$ at x = 0.

For $\beta > 0$, $h_f^- = h_a^+ = 0$ and, A-Eqs. (12a) (12b)

It is evident from Eqs. (12a, b) that for $\beta > 0$ either $\chi^+ = -2$ or $h_f^- = h^- = 0$.

Because $\chi^+ = -2$ (equivalent to $\mu^+ = -\mu_0$) is possible, a nontrivial solution to Eq. (12a) occurs when $\beta > 0$ and, A-Eq. 13.

This value of Ω is the normalized Damon-Eshbach frequency.

Because $\chi^- > 0$, Eq. (18b) has only the trivial solution $h_f = h_a = 0$ when $\beta > 0$.

Returning to the geometry of Figure 2, (with the bias taken along x_1), we find that the potential ψ within the ferrite region must contain both growing and decaying exponential terms. This

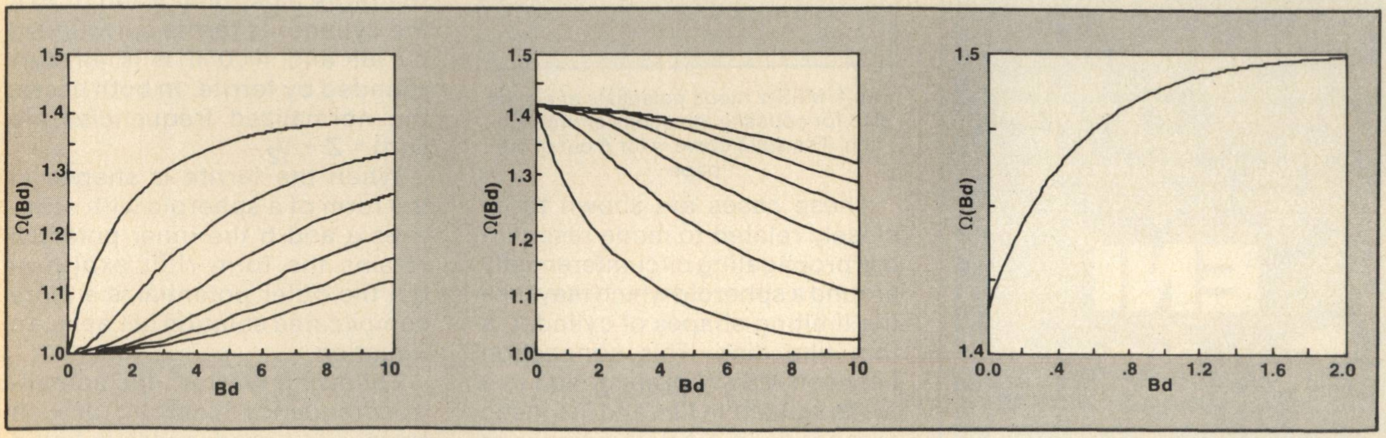
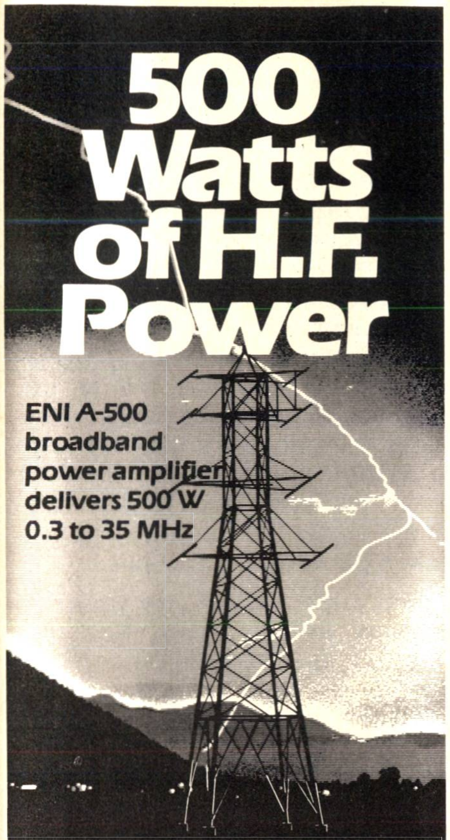


Fig. 3 Normalized MSW frequency dispersion for the geometry of Figure 2 when the dc magnetic bias is oriented parallel to (a) x_3-, (b) x_2- and (c) x_1- axis. The respective modes are termed MSFVW, MSBVW and MSSW.

[*Continued on page 86*]



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means that the polarization of $\bar{h}_f(x)$ will vary with x and hence χ^+, χ^- and $\exp|\beta|d$, $\exp|\beta|D$ will all enter the boundary conditions at $x = \pm d/2$ and $x = \pm(D+d/2)$. The results is surface wave propagation at the normalized frequency. A- Eqs. (14) (15)

The MSSW dispersion is shown plotted in Figure 3c for the case d/D = 0, Z = 1.

For $|\beta|d \ll 1$, the frequency is approximately $\Omega \approx \sqrt{Z(Z+1)}$ and the mode is uniformly distributed over the film thickness. For increasing $|\beta|d$, the frequency rises and the mode becomes increasingly localized on one of the two surfaces; the left most one for $\beta < 0$, the right most one for $\beta > 0$. When $|\beta|d \ll 1$, the mode frequency approaches the value $Z + 1/2$ as expected. For D = 0, the RF flux normal to the film is excluded ($\bar{h} = -\bar{m}$), and $\Omega = Z + 1$.

The MSSW mode potentials for the case $D = \infty$ are sketched in Figure 4 for opposite directions of propagation. The localization, which is stronger for increasing $|\beta|d$, occurs on the surface for which the mode has the largest positive-circularly-polarized component. Such field displacement nonreciprocity provides the underlying basis for many nonreciprocal devices, including Y-circulators.

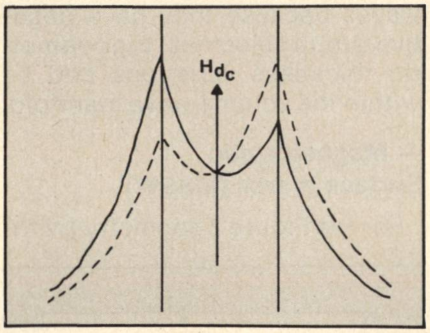

Fig. 4 MSSW mode potentials of a finite slab for opposite directions of propagation. The solid curve is for β out of the paper.

These cases are shown to be closely related to mode resonances propagating circumferentially around a spheroid which may take the limiting shapes of cylinder, a thin film disk. This connection between MSSW propagation in a rectangular thin film and magnetic resonance in a small cylinder or spere is emphasized because it provides a key to understanding the manner in which nonuniform bias fields can alter and hence control the frequency dispersion of MSW.[4]

— Width Variations and Edge Modes

The preceding analyses have assumed that the ferrite strip is thin ($d/w \ll 1$) and width variations negligible. In fact, such variations do occur and may be either trigonometric or exponential functions. If the former, we term them width modes; if the latter, edge modes.

Such variations are expected because if we reverse the aspect ratio and take $d/w \gg 1$, the FVW and SW geometries of Figure 2 effectively interchange (when $D = \infty$). For arbitrary d/w, it is expected that mode mixing [5,6] will occur even when $d/w \ll 1$.

— Similarity Between MSSW and Resonance Modes of a Spheroid

Returning to the semi-infinite slab geometry, we notice that if the surface is allowed to bend (with large radius of curvature) until it closes upon itself, a long cylinder is generated, either of ferrite surrounded by air (or vice versa). The x and y coordinates of the slab become equivalent to r and $s = r\phi$, and the value of β can be set equal to $\beta = n/r$ where m is an integer.

The magnetostatic potential within the cylinder, $\psi \simeq \exp(\int|\beta|dx - j|\beta|y)$ is $\psi \simeq \exp(\int \frac{|n|}{r} dr - jn/r\, r\phi) = r^{|n|} \exp(-jn\phi)$.

Because the inner and outer RF fields remain circularly-polarized, despite the curved path, modes therefore exist only when $n > 0$ if the cylinder is ferrite surrounded by air and $n < 0$ if it is air surrounded by ferrite. In both cases, the normalized frequencies are $\Omega(n) = Z + 1/2$.

When the ferrite is shaped in the form of a spheroid with semi-axes a and b the inner potential retains the form $r^{|m|} \exp(-jn\phi)$ but the outer potential is a more complicated solution of Laplace's Equation.

Following Walker[2], the normalized frequency spectrum for such surface like resonances is found to be A- Eqs. (16), (17) and (18).

For n = 1, the mode is the so-called uniform precession or Kittel resonance.

The uniform RF magnetization thus begins at the south poles on the surface of the spheroid and terminates on north poles on the opposite side. Such a pole distribution creates in turn an RF $\overline{h}$-field that is uniform inside the spheroid given by A- Eq. (19).
where N_t is the transverse demagnetizing factor of the spheroid (that is related to the axial demagnetizing factor, N_z by $2N_t + N_z = 1$).

Because $\overline{m}$ and $\overline{h}$ are positive-circularly-polarized vectors, $\overline{m} = \chi^+\overline{h}$ and nontrivial consistency is maintained with Eq. (19) only when $-\chi^+ N_t = 1$.
This occurs when $\Omega = Z + N_t$, which is consistent with Eq. (19) because $N(n=1) = N_t$; in addition $\lim_{n\to\infty} N(n) = 1/2$. Hence the mode spectrum starts with the n = 1 uniform precession (or Kittel resonance) and approaches the Damon-Eshbach frequency, Z + 1/2 as the magnetic potential concentrates ever closer to the surface due to its $r^{|n|}$ dependence. These limits are easily discerned in the case of a sphere (a = b) because then $N(n) = \frac{n}{2n+1}$. For b/a<<1 and moderate values of n, $N(n) \simeq b/2a\sqrt{\pi n}$ can be used as a resonable approximation for a thin disk.

— Nonuniform Bias

When the normalized bias is taken as in A- Eq. (20) with only weak to moderate gradients, the surface-like modes of a thin disk have normalized frequencies given by[4] A- Eq. (21)

Such modes develop small non-zero values of $\nabla\cdot\overline{h}$ throughout the volume which can change the state of polarization within the disk; in addition, both the mode frequency and the velocity with which the mode energy circulates are found to be altered.

For increased gradient strength, the volume divergence of certain modes can change so dramatically that selective localization or expulsion of the energy occurs. The sense of polarization can also actually reverse. In certain cases, the volume divergence of the RF magnetization can become very large at a certain interior radius. The magnetic pole distribution at this "virtual-surface" thus resemble that of a true surface and can serve to guide and localize the mode.

— Synthesized Dispersion Relations

If it is desired to create a spectrum with the resonance frequencies separated by pre-specified amounts or if one wishes to control the velocity of energy circulation $v^{E/\phi}$ of individual modes, the independent constants B_n can be adjusted and the required field $H_z(r)$ synthesized. In the former case Eq. (21) is used alone whereas in the latter, the relation $v^{E/\phi}(n) = r\partial\omega(n)/\partial n$ (analogous to the group velocity for a plane wave) is also employed.

It is useful to realize that $v^{E/\phi}$ can be forced to be independent of n over some range of n. This fact suggests that $H_z(x)$ can control MSFVW dispersion and that $\omega(\beta)$ can be approximately non-dispersive over a predetermined bandwidth.[7]

— Guided Magnetostatic Plane Waves

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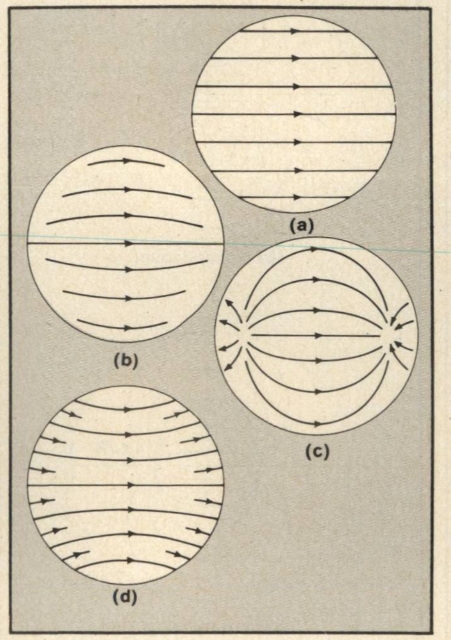

Fig. 5 RF contours, in the plane of a thin ferrite disk, that are associated with m = +1. In (a), the dc field gradient is zero; the Kittel uniform precession results; in (b) bias and mode energy are reduced near the disk rim; in (c) the reduction causes polarization reversal. In (d) increased bias and mode energy occurs near the rim.

[Continued on page 88]

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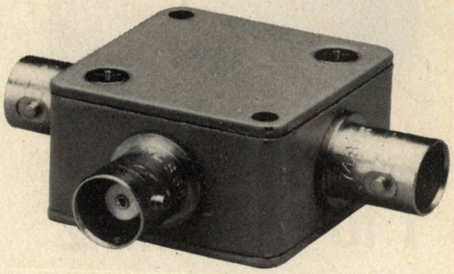

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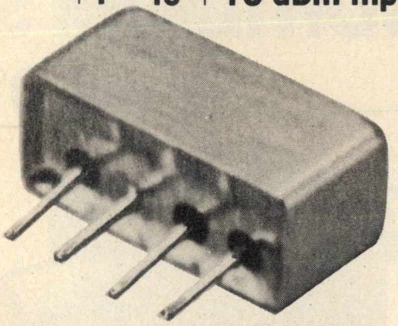

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F3	−50	−40
200-600 MHz F1	−25	−20
F3	−40	−30
600-1000 MHz F1	−20	−15
F3	−30	−25

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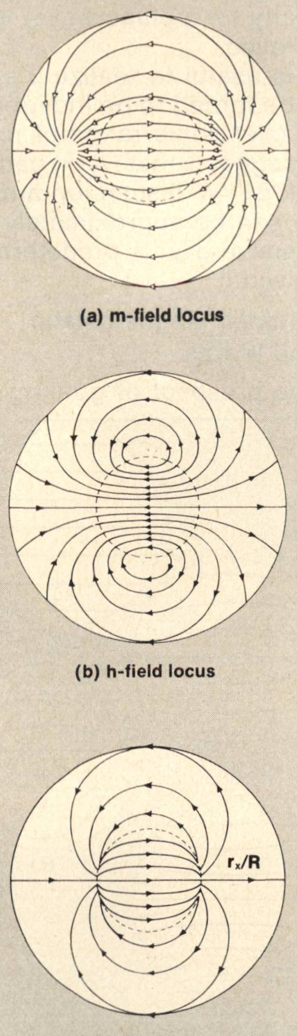
netostatic waves is of considerable interest because of possible device applications.[8] Such guided waves might be used to increase the delay time realizable on a given size sample by meandering the path, or to make a resonator by guiding the waves along a closed loop. In addition, controlling the coupling between the adjacent waveguides is important for signal routing devices such as directional couplers and diplexers.[9]

For these reasons, the analysis and synthesis of magnetostatic surface and volume waves (MSSW and MSVW) when the dc bias is nonuniform takes on special significance. For example, guiding the MSSW around the bend is complicated due to mode conversion to backward volume waves, but it has proved possible to overcome this difficulty by employing gradients that arise from a change in the direction of the bias field.[8]



(a) m-field locus

(b) h-field locus

(c) b field locus

Fig. 6 Locus of field lines for the "virtual-surface" mode $\Omega(m = 1) = .655$ when $Z = .3 + 1.095\,(r/R)^4$. The rf m-field is shown in (a) the h-field in (b) and the b-field in (c). All patterns rotate at the normalized frequency Ω. The "virtual-surface" (shown dotted) occurs at $r_x/R = .385$.

The geometry used to investigate guiding a surface wave through a turn of 160° is shown in Figure 5. The 4.5 μmeter thick YIG film is in the shape of a half disk one inch in diameter.

The inner radius of the channel of ~ .4 cm and the outer radius is ~ .9 cm resulting in a channel width of ~ .5 cm, and a mean radius of ~ .6 cm. The mean circumferential path between antennas is ~ 1.8 cm. The transmission characteristics are shown in Fig. 6.

Since the YIG exists beneath the mu-metal as well as in the channel, other propagation paths than the channel are conceivable, though unlikely due to the change in field direction. In order to verify that the signal was propagating around the curved path, the wave was probed using a small induction loop at several points along the path. In addition to verifying the presence of the wave at various points along the path, the loop was used to determine the local direction of the k vector. This was done by making use of the fact that the loop is most sensitive to k vectors directed normal to its plane; hence the direction of k can be approximately determined by rotating the loop. The results of these measurements, while not all precisely circumferentially directed, strongly support the waveguide interpretation. The curved path experiment is believed to be the first demonstration of the ability to guide MSSW's.

SUMMARY

Uniform MSW requires the RF magnetic permeability to be negative. This is possible over a band of frequencies that constitutes the volume wave manifold. Transverse boundary conditions and bias orientation determine whether the frequency dispersion has forward or backward wave character.

Nonuniform MSW bound to an air-ferrite interface can propagate when the bias is in-plane and transverse. Because the RF potential is then circularly-polarized, the MSW mode requires that $\mu^+ = -\mu_o$. Such a surface wave can propagate in only one direction because the sense of polarization reverses for the opposite direction and μ^- cannot take on negative values.

Nonuniform magnetic bias can alter the character of MSW modes, affecting the frequency dispersion and RF energy distribution. Such control should allow one to synthesize MSW devices for novel microwave signal processing applications.

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Frederic R. Morgenthaler *graduated from the Massachusetts Institute of Technology with BS and MS degrees in Electrical Engineering in June 1956 and the Ph.D. degree in June 1960.*

Dr. Morgenthaler is currently Professor of Electrical Engineering at M.I.T. During the academic years 1960 to 1962, he was a Ford Foundation Post Doctoral Fellow in Engineering and an Assistant Professor in the M.I.T. Department of Electrical Engineering. In 1965 he was promoted to the rank of Associate Professor and in 1968 to the rank of Professor.

Dr. Morgenthaler is a member of the Research Laboratory of Electronics, the Center for Materials Science and Engineering, and Director of the Microwave and Quantum Magnetics Group. He also holds an appointment to the Harvard-M.I.T. Division of Health Sciences and Technology. His research, and graduate teaching, has centered on the fields of microwave magnetics and ultrasonics, the electrodynamics of waves and media, and microwave signal processing using coherent wave states in solids. In addition, Dr. Morgenthaler participates in an interdisciplinary course in medical ultrasonics. He has also taught the undergraduate Electrical Engineering core curriculum subjects in electromagnetic field theory, circuit theory, and semiconductor electronics. He was in charge of the field theory course from 1963-1968 and lecturer in charge of the transistor course from 1973-1978.

Dr. Morgenthaler serves as a consultant to the U.S. Government as well as to private industry. He is a Fellow of the Institute of Electrical and Electronics Engineers and a member of the American Physical Society, the American Association for the Advancement of Science; and of the Sigma Xi, Tau Beta Phi and Eta Kappa Nu honorary professional societies. He is the author of numerous scientific publications and papers presented at technical conferences and has been granted approximately one dozen patents with several others pending.

[*Continued on page 90*]

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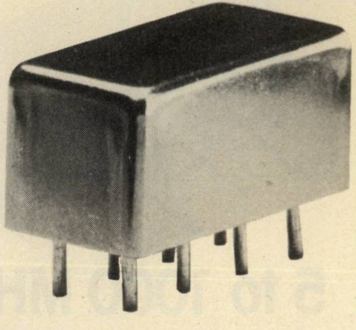

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100-1000 MHz	0.7	1.2
ISOLATION, dB	25dB	TYP.
AMPLITUDE UNBAL.	0.2	TYP.
PHASE UNBAL.	2°	TYP.
IMPEDANCE	50 ohms.	

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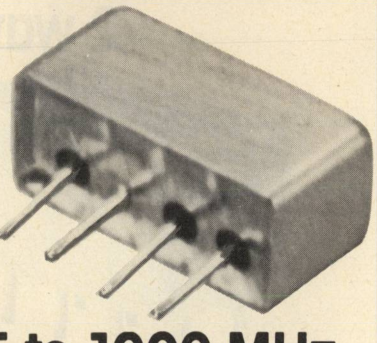

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LO-IF	45	40
LO-RF	40	30
LO-IF	35	25
LO-RF	30	20
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$$\frac{\partial \bar{M}}{\partial t} = \gamma\mu_o \; \bar{M}\times\bar{H}^E \quad (1) \qquad \frac{\partial \bar{m}}{\partial t} = \bar{i}_z \times (\omega_z\bar{m} - \omega_M\bar{h}) \quad (2)$$

$$\bar{m} = m_o[\bar{i}_x \cos(\omega_z t+\phi) + \bar{i}_y \sin(\omega_z t+\phi)] \quad (3)$$

$$\bar{h} = \frac{\bar{\gamma}(\bar{\gamma}\cdot\bar{m}) + \omega^2\epsilon\mu_o\bar{m}}{-(\bar{\gamma}\cdot\bar{\gamma}+\omega^2\epsilon\mu_o)} \quad (4)$$

$$(1+\chi)\left(\frac{\partial^2\psi}{\partial x^2} + \frac{\partial^2\psi}{\partial y^2}\right) + \frac{\partial^2\psi}{\partial z^2} = 0 \quad (5)$$

$$\Omega^2 = Z^2 + Z\frac{\gamma_x^2+\gamma_y^2}{\gamma_x^2+\gamma_y^2+\gamma_z^2} \quad (6) \qquad \Omega^2 = Z^2 + Z\frac{\beta^2}{\beta^2+k_z^2} \quad (7)$$

$$\beta \tanh \beta D = k_z \begin{cases} \tan \frac{k_z d}{2} & \text{(even)} \\ -\cot \frac{k_2 d}{2} & \text{(odd)} \end{cases} \quad (8) \qquad \Omega^2 = Z^2 + Z\frac{k_y^2}{k_y^2+\beta^2} \quad (9)$$

$$\beta\tanh \beta D = k_y \begin{cases} -\cot\left(\frac{k_y d}{2}\right) & \text{(even)} \\ \tan\left(\frac{k_y d}{2}\right) & \text{(odd)} \end{cases} \quad (10)$$

$$\psi = \begin{cases} C_f e^{|\beta| x} e^{-j\beta y} & x \le 0 \\ C_a e^{-|\beta| x} e^{-j\beta y} & x \ge 0 \end{cases} \quad (11)$$

For $\beta > 0$, $h_f^- = h_a^+ = 0$ and

$$(1+\chi^+)(h_f^+)_{x=0} = (h_a^-)_{x=0} = -(h_f^+)_{x=0} \quad (12a)$$

whereas for $\beta < 0$, $h_f^+ = h_a^- = 0$ and

$$(1+\chi^-)(h_f^-)_{x=0} = (h_a^+)_{x=0} = -(h_f^-)_{x=0} \quad (12b)$$

$$\Omega = Z + \frac{1}{2} \quad (13)$$

$$\Omega^2 = Z^2 + \frac{2Z(1+\tanh|\beta|D) + 1 - [2Z(1-\tanh|\beta|D)+1]e^{-2|\beta|d}}{(1+\tanh|\beta|D)^2 - (1-\tanh|\beta|D)^2 e^{-2|\beta|d}} \quad (14)$$

For $D = \infty$, Eq. (14) reduces to

$$\Omega = \sqrt{\left(Z+\frac{1}{2}\right)^2 - \frac{1}{4}e^{-2|\beta|d}} \quad (15)$$

$$\Omega(n) = Z + N(n) \quad (16)$$

where

$$N(n) = \frac{b}{a}\frac{\left(\frac{a}{c}\right)^{n+1}}{(n-1)!2^n}F_n^{(n)}\left(\frac{a}{c}\right) \quad (17)$$

and

$$F_n^{(n)}(\zeta) = (2n-1)!!\,\zeta^n \sin^{-1}\left(\frac{1}{\zeta}\right) - \frac{\sqrt{\zeta^2-1}}{\zeta^n}[(2n-2)!!+(2n-1)(2n-4)!!\zeta^2 + (2n-1)(2n-3)(2n-6)!!\,\zeta^4 + ..+(2n-1)!!\,\zeta^{2n-2}]$$

with $c = \sqrt{a^2-b^2}$ and $p!! = p(p-2)(p-4)...$ (18)

$$\bar{h} = -N_t\,\bar{m} \quad (19) \qquad H_z(r)/M = Z(r) = A + \sum_{n=1}^{N} B_n r^{2n} \quad (20)$$

$$\Omega(n) = A+N(n) + \frac{1}{1+\frac{3/2}{n}}\left(B_1 a^2 + \frac{1}{1+\frac{3/2}{n+1}}\left(B_2 a^4 + \frac{1}{1+\frac{3/2}{n+2}}\left(B_3 a^6 + \ldots\right.\right.\right. \quad (21)$$

■

Hybrid Saw Oscillators

S. Neylon
Marconi Electronic Devices Ltd.

Introduction

Surface Acoustic Wave Oscillators (SAWOs) have been competing as alternative IF frequency sources to bulk wave crystal oscillators and cavity oscillators for several years and are currently used in numerous military and commercial applications.

As a planar element the SAW chip readily lends itself to integration into standard hybrid circuits. This paper reviews the fabrication of hybrid SAWOs and presents results based on two such oscillators. This type of device offers a number of advantages to microwave systems engineers:

- Fundamental frequencies 20-1500MHz
- Good frequency stability
- Favourable close to carrier and good wideband noise levels.
- Ruggedness
- FM capability
- Small size and low mass.

Devices have been fabricated from 50-1599 MHz for a wide variety of applications. The high fundamental frequency allows low multiplication factors, thereby achieving stable microwave frequencies while simultaneously simplifying harmonic filtering. Oscillators have also been specifically designed to exploit the above advantages. Their ruggedness and small size make SAWOs ideal for use in extreme ECM environments, whilst their good wideband noise level facilitates use in phase locked loop subsystems. In addition frequency agility has been demonstrated by multi-mode device operation or by switching between a number of individual oscillators.

Design and Operation

The basic principles in the design and operation of SAWOs have already been well documented.[1,2] In its simplest form the SAW element acts as the frequency selective device in a feedback loop network (Figure 1). The SAW element can, however, be designed to perform in one of two distinct modes; as a delay line or as a resonator. In either form, single mode of operation is ensured by designing into the transducer structures a unique phase/frequency relationship within the passband. The phase slope $d\Phi/df$ defines the Q of the device.

$$Q = \frac{f_0}{2} \frac{d\Phi}{df}$$

In many applications, it is this Q factor which determines whether a SAW resonator or delay line is used.

SAW resonators offer a high Q capability ($\simeq$ 10,000) with an associated good close to carrier spectral purity and frequency stability. However, the fine geometries involved in defining a high frequency SAW resonator presently limit their higher fundamental frequency of operation to 1200 MHz.

Where lower Q's or wide modulation bandwidths are required, a SAW delay line is most commonly used. Although high Q's are similarly obtainable using SAW delay lines, the increased substrate size relative to resonators, can be a limiting factor. The increased insertion loss of the delay line approach also necessitates additional amplification in the oscillator loop. Linewidth definition is, however, less demanding and fundamental frequencies up to 1500MHz are currently available.

The oscillator shown in Figure 1 consists of amplifiers, phase shifter, matching networks, resistive power splitter and the SAW element. Oscillation will occur if the loop gain is unity and at a frequency where the loop phase is an integral multiple of 2π. Provided the loop phase is within the linear phase/frequency range of the SAW element, introduction of a phase shift in the circuitry will cause a shift in the frequency of oscillation by:

$$\Delta f = \frac{f_0}{2Q} \cdot \Delta\Phi$$

An optimal value of the Q can therefore be chosen for any specific application. For fixed frequency operation, a high Q is desirable to minimize the effects of circuit phase changes and hence the use of a SAW resonator. Where frequency/phase modulation or voltage control oscillators are required, suitable electronic phase shifters can be built into

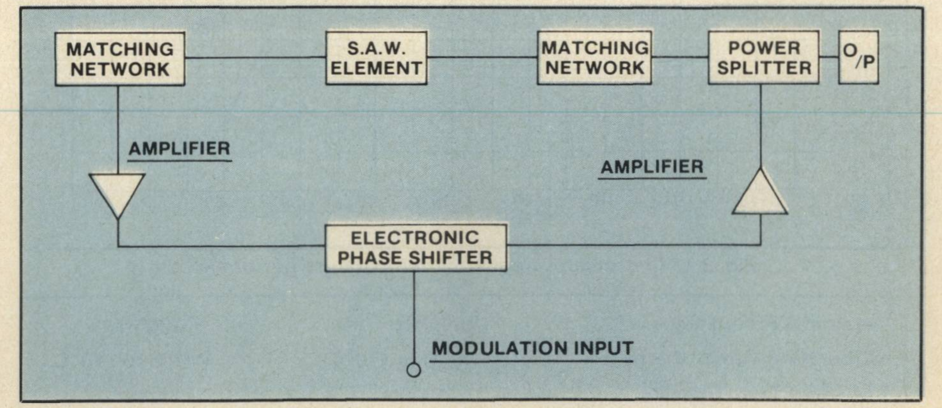

Fig. 1 Block diagram of typical SAW oscillator.

the loop in conjunction with a SAW delay line.

The simplest method of generating acoustic waves is to use piezoelectric substrates. The Long Term Stability (LTS) and good temperature coefficient of quartz recommends it for most applications, and, being anisotropic, specific orientations or 'cuts' can be selected to optimize the SAWO temperature performance.

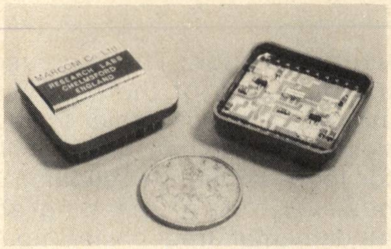

Fig. 2 A fixed UHF oscillator.

Hybrid Oscillators

The miniaturization of circuits using thick film technology has been well established in recent years and the availability of the SAW element as a planar chip lends itself easily to integration using this approach. Advantages in producing SAWO's in this way include high reliability in extreme environments, small size and low mass. The ruggedness of the SAW chip complements the rigid construction techniques of the thick film circuits, thereby maximizing the device immunity to shock and vibration. In addition, inductors, capacitors and transistors can be added in chip form or as encapsulated components and mounting techniques adjusted to optimize the long term stability and ruggedness of the device. The close control over circuit parameters achieved in thick film circuits minimizes the need for alignment and coupled with the use of automatic test equipment, allows costs to be reduced when compared with standard printed circuit assemblies.

A range of hybrid circuits over the frequency range 200-1200MHz have been developed with the flexibility to incorporate either SAW resonator or delay line chips. Two such hybrid oscillators are presented, which exemplify these advantages.

A fixed source in the mid UHF range was developed for an application requiring high frequency stability, and is shown in Figure 2. A SAW resonator was chosen as the frequency selective element for the obvious advantages of small size, low insertion loss and high Q, thereby minimizing the distortions from amplifiers and circuitry.

The thick film circuit has the flexibility to accommodate either SAW delay lines or SAW resonators operating from 200-700MHz. Two stages of amplification are available in the feedback loop lest a SAW delay line is used, but in this case the low insertion loss of the SAW resonator dictates the use of only one of these stages. An output buffer amplifier has also been incorporated into the design layout to minimize load pulling and improve the output match. Provision in the design has also been made for several circuit options, including a varactor phase shifter.

A low cost UHF type transistor in SOT-23 encapsulation were used in the loop amplification and output buffer stages because of electrical suitability and ease of handling. It was envisaged however that the plastic encapsulant might prove detrimental in achieving long term stability and equivalent chip devices are envisaged for use in future production devices.

Matching networks included flat spiral inductors on thin alumina substrates; standard chip capacitors, and laser or abrasion trimmed resistors.

Several devices were built to this design and evaluated. Frequency/temperature performance results (Figure 3) over the range -10 to +60°C indicated a turnover point at 15°C. This factor is determined largely by the 'cut' of quartz used and suitable orientation adjustments could yield a

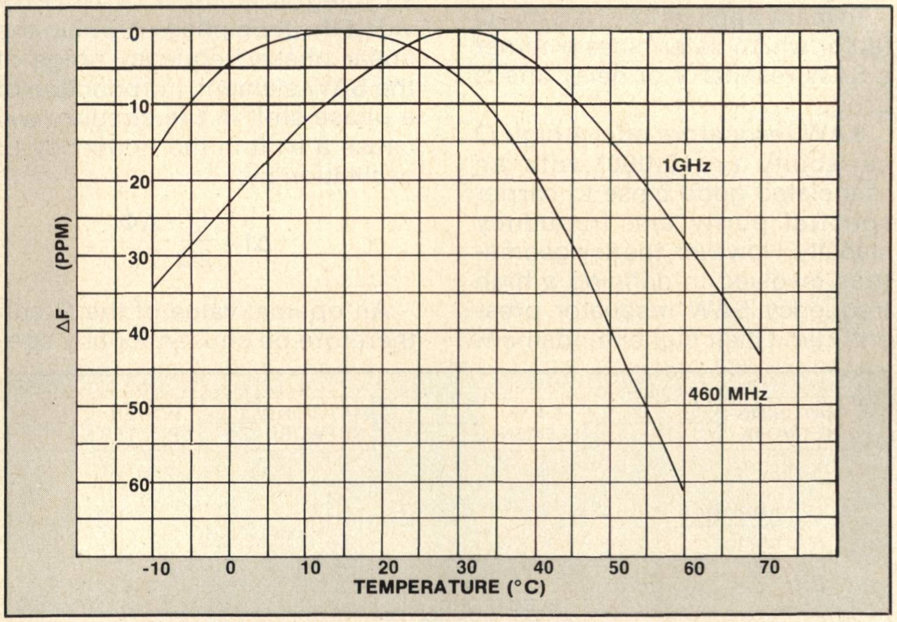

Fig. 3 UHF oscillator frequency/temperature performance.

Nominal Frequency	460MHz	1020MHz
Operating Temp. Range	-10°C to +70°C	-10°C to +70°C
Medium Term Stability	±40ppm	±25ppm
Long Term Stability	5ppm	5ppm
Power Output	6dBm	6dBm
Power Supply	12V <40mA	12V <40mA
Harmonic Content	-20dBc	-20dBc
Shock	3000g	3000g
Vibration	0.3g	0.3g
Size	30x30x5mm	30x30x5mm

Fig. 4 Target production specification.

more centralized turnover point. On the basis of these results a target production specification has been defined as shown in Figure 4.

A hybrid SAWO (Figure 5) operating at a nominal frequency of 1020MHz.

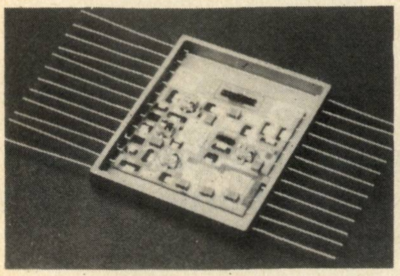

Fig. 5 A hybrid 1020 MHz SAWO.

Thick film circuitry is once again designed for flexibility. Covering the frequency range 500-1200MHz, all the options of the previous circuit have been included; SAW resonator or delay line, two stages of loop amplification; output buffer; varactor diode phase shifter or pin diode switching. In addition a resistive power splitter has been incorporated to isolate output mismatch. Once again conventional hybrid matching components have been used and the option to use encapsulated or chip transistors and varactor

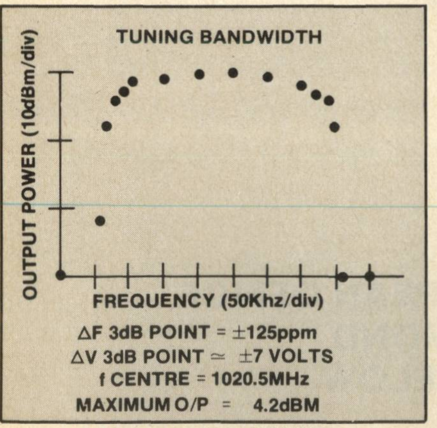

Fig. 6 Varactor-tuned SAWO tuning bandwidth.

diodes is available.

This particular device is based on a SAW delay line at 1020MHz. Both loop amplification stages are used in addition to the output amplifier. Protection against low frequency oscillation has also been designed into the circuitry.

Using this as a basic device, two variants have been built. One incorporates a varactor diode, giving ≃ 200kHz FM tuning capability over the voltage range 0-12v as illustrated in Figure 6. The other replaces the varactor diode with a T.T.L. compatible PIN switching network.

A number of devices, demonstrating the different options, have

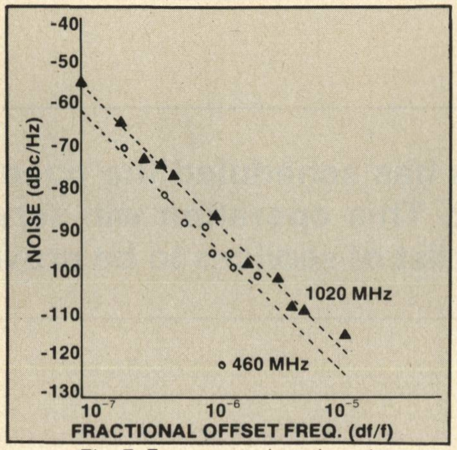

Fig. 7 Frequency domain noise measurements.

been built and assessed. Based on these results a target specification has been formulated. See Figure 4. Noise measurements conform closely to standard SAWO spectral purity theory[3] (see Figure 7).

Conclusions

A wide range of hybrid SAWOs have been built for a variety of applications in the 200-1300 MHz range. The ruggedness, low cost and flexibility of oscillators has been proved using this approach, and now argues strongly for their further use in military and communication systems.

Acknowledgements

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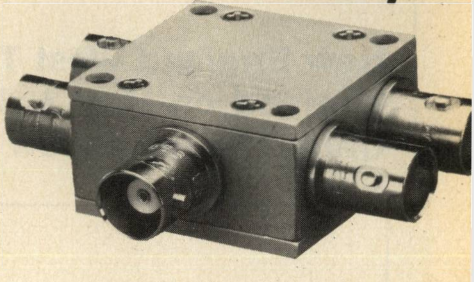

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AMPLITUDE UNBAL., dB	0.1	0.2
PHASE UNBAL. (degrees)	1.0	4.0
ISOLATION, dB	TYP.	MIN.
(adjacent ports)	23	20
ISOLATION, db (opposite ports)	23	20
IMPEDANCE	50 ohms.	

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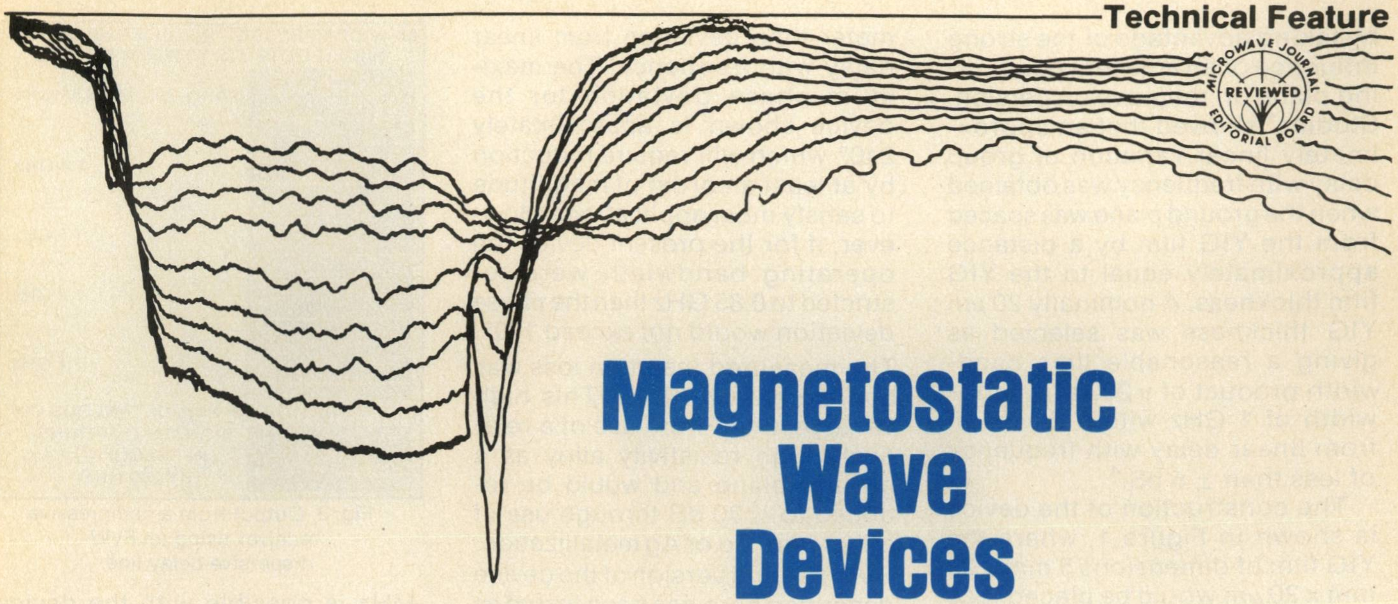

Magnetostatic Wave Devices

J.D. Adam, Michael R. Daniel and T.W. O'Keeffe
Westinghouse R&D Center
Pittsburgh, PA

Introduction

Magnetostatic wave (MSW) devices, based on epitaxial YIG are potentially useful for signal processing at microwave signal frequencies or at a broadband microwave IF. MSWs are attractive because of their low propagation losses of less than 30 dB/μS at microwave frequencies and their planar geometry and bandwidths, of the order of 1 GHz, are compatible with presently available semiconductor amplifiers or signal sources. The objective here is to describe several types of devices which have been recently demonstrated and to outline potential applications of a few of them. The devices are a dispersive delay line, a programmable tapped delay line and a 10-channel filter bank, all operating in X-band.[1] Equations describing FVW propagation and transduction in structures consisting of a YIG film spaced from a ground plane have been described elsewhere[2, 3] and will not be discussed here.

The forward volume wave (FVW) propagation was used for the delay devices because of the wide bandwidths (> 1 GHz) available with this mode at X-band which would be impossible to achieve with a simple surface wave device.

Dispersive Delay Line

The desired delay versus frequency characteristics of the dispersive delay line were obtained

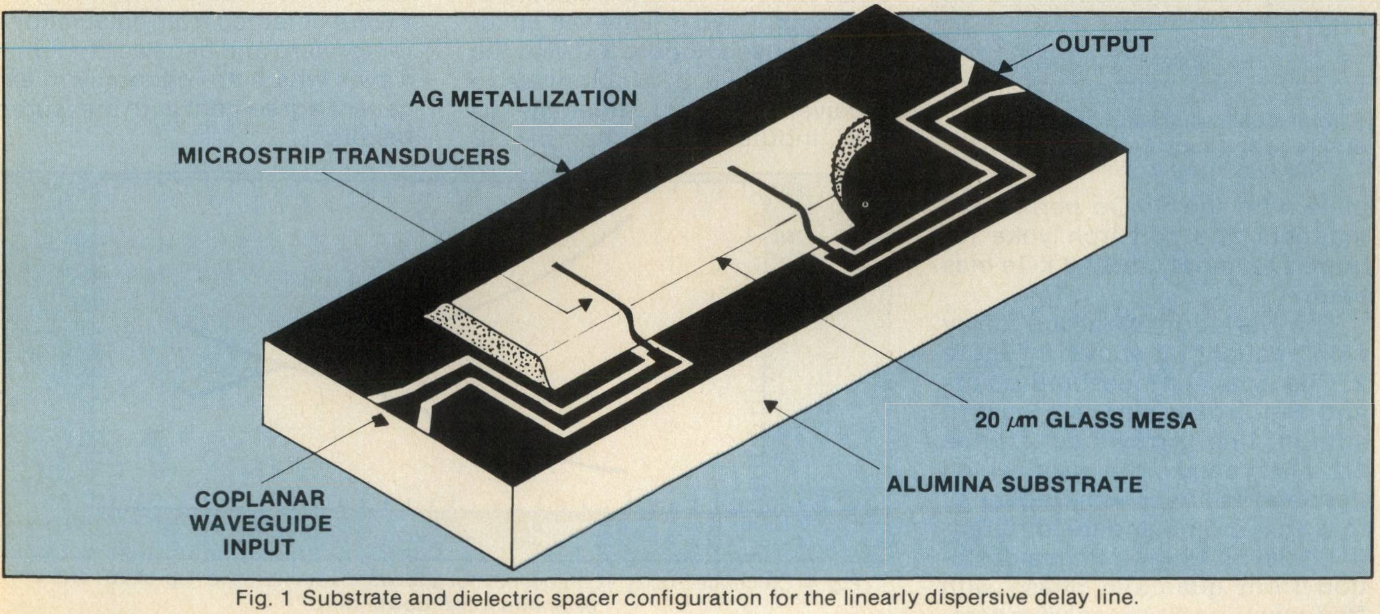

Fig. 1 Substrate and dielectric spacer configuration for the linearly dispersive delay line.

by taking advantage of the strong influence of an adjacent conducting plane on the wave dispersion. Studies[4] showed that an approximately linear variation of group delay with frequency was obtained when the ground plane was spaced from the YIG film by a distance approximately equal to the YIG film thickness. A nominally 20 μm YIG thickness was selected as giving a reasonable time bandwidth product of ν 200 at a bandwidth of 1 GHz with a deviation from linear delay with frequency of less than $\pm$ 5 nS.[4]

The construction of the device is shown in Figure 1, where the YIG film of dimensions 5 mm x 25 mm x 20 μm would be placed face down on the glass mesa so as to contact the microstrip transducers. The glass mesa was deposited on the gold ground plane by sputtering. The gold microstrip transducers were 5 mm long, 50 μm wide and 5 μm thick and were open circuited at one end with the opposite end connected to coplanar waveguide of 50Ω characteristic impedance. The packaged device is shown in Figure 2 complete with the SmCo permanent magnets and soft iron yoke required to supply the 4.8 kOe bias field.

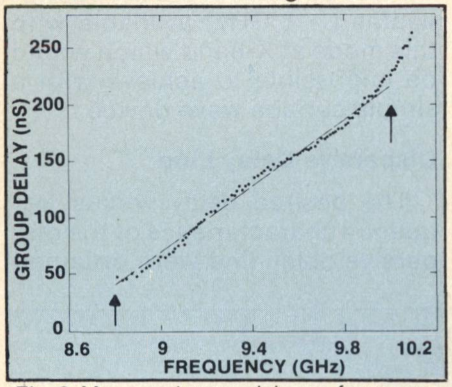

Fig. 2 Measured group delay vs. frequency for the delay line shown in Figure 1; arrows denote bandwidth over which the delay is linear to within $\pm$ 5 nS.

The variation in group delay with frequency is shown in Figure 2. The dots are measured points and the broken line is a best fit straight line. Arrows denote the 1.2 GHz bandwidth over which the delay is linear with frequency to within $\pm$ 5 nS and the differential delay is 190 nS. Phase deviation from quadratic phase with frequency is a more sensitive parameter than deviation from linear delay with frequency. The maximum phase deviation for the device shown is approximately 240° which will require reduction by at least an order of magnitude to satisfy most applications. However, if for the present device the operating bandwidth were restricted to 0.85 GHz then the phase deviation would not exceed 100°.

The measured insertion loss was approximately 40 dB. This high value resulted from use of a relatively high resistivity alloy as a ground plane and would be reduced to $<$ 30 dB through use of either pure Au or Ag metallization.

An S-band version of the device described here has been tested as the dispersive delay element in a microscan receiver where the microwave operating frequency and 1 GHz bandwidths of the magnetostatic wave device offer improvement in performance in certain situations compared with systems using surface acoustic wave devices. In a simple compressive receiver an incoming signal is mixed with a linearly frequency swept local oscillator. The mixer output is fed to a linearly dispersive delay line whose delay versus frequency slope is equal but opposite in sign to the local oscillator sweep. The output of the delay line is then essentially the Fourier transform of the input signal. This technique is applicable to rapid wideband spectrum analysis with 100% probability of intercept if two interleaved channels are used. Figure 3 shows the output from an MSW compressive receiver with two simultaneous C.W. inputs. A resolution of ~12 MHz is possible with the device used.

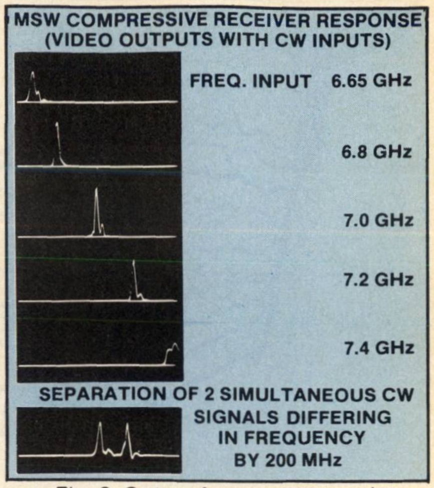

Fig. 3 Output from a compressive receiver using an FVW dispersive delay line.

Another important potential application of MSW delay lines is as the delay element in phased array antennas. Each element or perhaps sub-array of the phased array antenna would be fed from a constant but electronically adjustable delay line so as to give an electronically steered beam with wide instantaneous bandwidth. One approach to achieving variable delay is to use two delay lines, with the characteristic shown in Figure 4 in cascade so as to give a total delay which was constant with frequency.[5] The total delay could then be varied by changing the magnetic bias field on one of the delay lines so as to move its delay characteristic along the frequency axis. The curves shown in Figure 4 were obtained from calculations for forward and backward volume waves which are generally in excellent agreement with measured results.

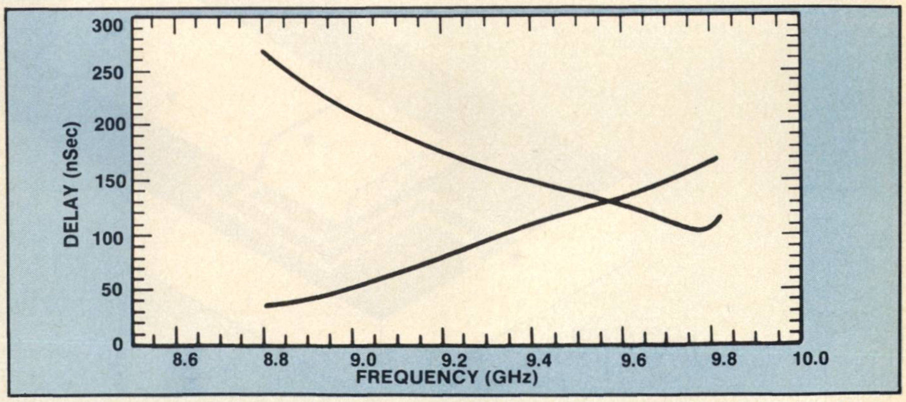

Fig. 4 Calculated delay vs. frequency for a forward volume wave "up-chirp" and a backward wave "down-chirp."

An example of the calculated delay variation, which can be obtained using presently available techniques, is shown in Figure 5 where the bandwidth of interest was limited so as to include the most linear portions of each delay characteristic in Figure 4. Here the phase deviation from a linear phase variation with frequency is plotted for three different values of bias field applied to the forward volume wave device. The bias was increased in 30 Oe steps between curves a, b, and c resulting in a delay increment of 10 nS between the curves. The tolerances on the phase deviation for a phased array antenna application are severe, typically less than 2 or 3°. In Figure 5, ± 18° tolerances were obtained over a 100 MHz bandwidth for an adjustable delay range of 10 nS between curves b and c. However, if the calculations were performed for parameters which limited the absolute delay to the range 5 to 15 nS instead of 250 to 260 nS (Figure 5), then a phase error of only ± 5° results over a bandwidth of 700 MHz. In phased array applications adjustable delays of 10 nS are adequate for beam steering antennas of aperture 12 feet.

Both broadband dispersive delay and adjustable delay are two important functions which cannot readily be achieved by other techniques at microwave frequencies, except by means of MSW technology.

Programmable Tapped Delay Line

The device discussed in this section is a programmable tapped delay line[6] which can be used to generate and correlate a phase coded pulse sequence. The 4-tap device is shown in Figure 6. The YIG film is the dark strip towards the bottom of the photograph and is 7.1 μm thick, 2 cm long and 1 mm wide. The input transducer is at the bottom center and the four output taps are arranged two on either side of the input so that there is a 1 mm path length increment between the input transducer and successive output taps. The taps are connected via 50Ω microstrip to PIN diode, 0 or π, phase shifters and then combined together. The microstrip circuit was fabricated on a 10 mil thick alumina substrate, 1" square. Switching and bias inputs for the PIN switches can be seen on both sides of the box. When completely assembled, the field produced by the permanent bias magnets was normal to the film. The insertion loss from the input to any output was typically 22 dB.

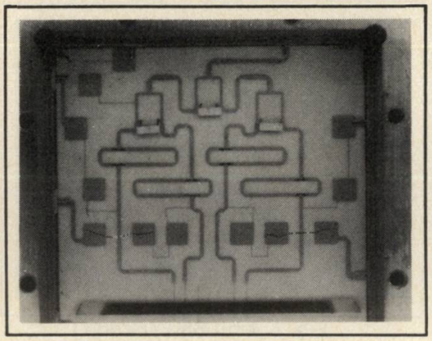

Fig. 6 Programmable 4-tap delay line.

The tapped delay line was tested by using the same device to generate and then correlate a 4-bit Barker code. A 9 GHz pulse of length 20 nsec was applied to the input transducer resulting in a phase coded train of four pulses from the output of the tapped

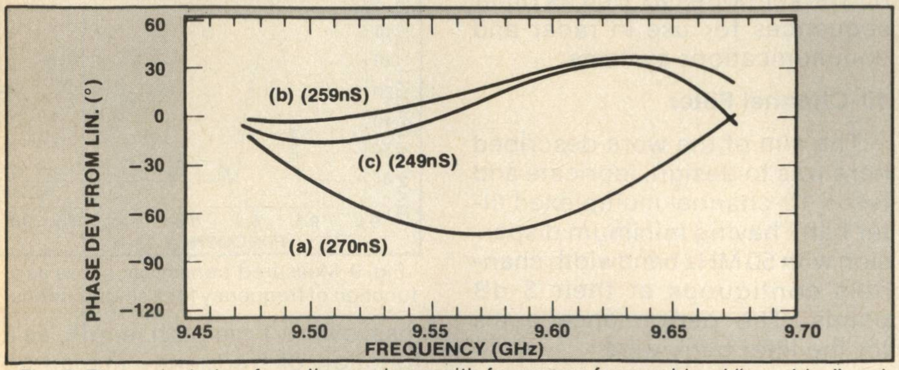

Fig. 5 Phase deviation from linear phase with frequency for combined "up-chirp" and "down-chirp" delay lines. Bias field applied to "up-chirp" device was increased in 30 Oe steps between (a), (b) and (c). The corresponding constant delay values are shown.

[Continued on page 98]

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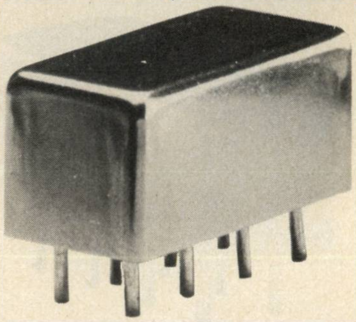

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FREQUENCY RANGE, (MHz)			
LO, RF	.05 - 2000		
IF	.05 - 500		
CONVERSION LOSS, dB		TYP.	MAX.
One octave from band edge		6.0	7.5
Total range		7.0	9.0
ISOLATION, dB		TYP.	MIN.
.05-.5	LO-RF	25	20
	LO-IF	25	20
.5-1000	LO-RF	40	30
	LO-IF	40	30
1000-2000	LO-RF	30	20
	LO-IF	25	15

Signal 1 dB Compression level +3dBm

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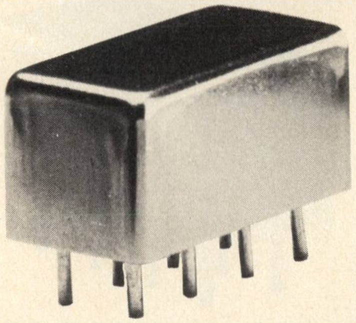

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SRA-6 SPECIFICATIONS

FREQUENCY RANGE, (MHz)			
LO, RF	3KHz - 100		
IF	DC-100		
CONVERSION LOSS, dB		TYP.	MAX.
One octave from band edge		5.5	7.5
Total range		6.5	8.5
ISOLATION, dB		TYP.	MIN.
.003-.03	LO-RF	60	50
	LO-IF	60	45
.03-50	LO-RF	45	30
	LO-IF	40	25
50-100	LO-RF	35	25
	LO-IF	30	20

Signal 1 dB Compression level +1dBm

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delay line. The phase coded pulses were delayed for 100 nsec in a constant MSW delay line and during this time the phase of the four taps was changed by an exerciser so as to represent a time reversed version of the generated pulse train. The pulse train was then amplified and reintroduced to the input of the tapped delay line via a directional coupler. The tapped delay line acted as a matched filter and performed a correlation on the four coded pulses. The sequence of pulses observed at the output to the tapped delay line is shown in Figure 7. Feedthrough from the input

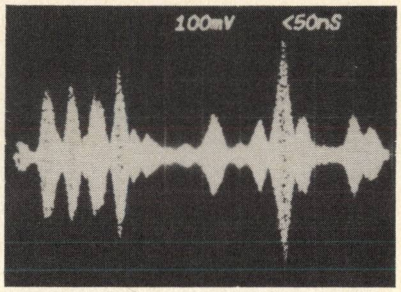

Fig. 7 4-bit Barker code generation and correlation.

pulse is seen on the extreme left followed by the four generated pulses which are coded with the π, π, o, π Barker code. The correlation peak is seen towards the right of the photo. For this 4-bit Barker code the correlation is a 7 pulse sequence with normalized amplitudes 1, 0, 1, 4, 1, 0, 1. The experimental results are in good agreement with this although some extra undesired pulses are evident which are due to reflections and higher order width modes. The techniques described are capable of extension to the generation and correlation of 13-bit Barker codes as well as other sequences for use in radar and communications systems.

10-Channel Filter

The aim of the work described here was to design, fabricate and test a 10-channel multiplexed filter bank having minimum dispersion with 50 MHz bandwidth channels contiguous at their 3 dB points. The performance goals for the filter bank were:

Center frequency	9.0 GHz
Number of channels	10
Channel 3 dB bandwidth	50 MHz
Out of band rejection	55 dB
50 dB bandwidth	100 MHz
Multiplexed insertion loss	20 dB
Bandpass ripple	1 dB

The filter band was designed as 10 narrow band delay lines, each fed from a common input transducer but with separate output transducers, as shown in Figure 8. A bias field gradient was applied along the device so that each delay line experienced a different bias field and hence has a different center frequency. The transducers were formed from 50Ω impedance microstrip line on 0.635 mm thick alumina and were open circuited at one end. The position of the delay lines along the input transducer was such that each was close to (N-½) electromagnetic half wavelengths from the open circuited end at its center frequency.

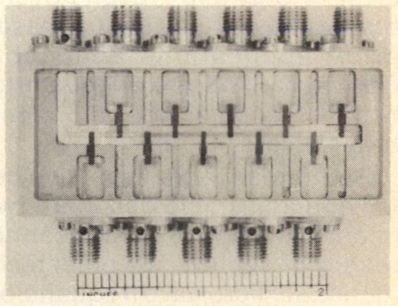

Fig. 8 Interior of 10-channel filter bank.

Narrow band pass characteristics were obtained by use of wide transducers (0.635 mm) spaced 160 μm from the YIG surface. Figure 9 shows the transmission

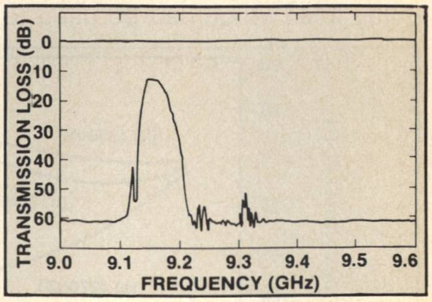

Fig. 9 Measured transmission loss as a function of frequency for a single channel.

loss, over a 1 cm path length, as a function of frequency measured on an 18.5 μm thick YIG film. This YIG film had 18 evaporated aluminum strips 264 Å thick, to attenuate higher order width modes, i.e. modes with transverse wavenumbers $k = \pi/W$, $3\pi/W$, $5\pi/W$ where W is the width of the YIG strip, and reflecting ends.[1] The spacing from the ends of the film to the edge of the transducer was critical and it was found that a spacing of 0.31 mm gave a reasonably symmetrical passband shape. The undesired out of band responses are approximately 40 dB down on the desired passband.

The measured transmission and return loss for all 10 channels as a function of frequency is shown in Figure 10. In these results the out of band responses are higher than desired. The non-uniform spacing of the passband center frequencies is due to the nonlinear variation in bias field along the device length. In addition, variations in the spacing of the ends of the YIG from the transducers resulted in changes in the passband shapes. It has been shown that a degree of control of the passband shape can be obtained with simple single element microstrip transducers. Further control may be possible with relatively simple multi-element transducers.

Future Projections

At present the most fruitful approach to wider bandwidth and closer tolerances on phase appears to be through the use of geometrical factors such as ground planes and double films to modify the delay with frequency so as to approximate the desired characteristic followed by careful application of SAW analogs or other techniques to reduce the phase error to acceptable levels.

Acknowledgments

The authors thank Dr. P. R. Emtage for his invaluable contributions to the development of the devices described here and also C. Nothnick for making results on the compressive receiver available. The work reported was supported in part by the U.S. Air Force Avionics Laboratory under Contract No. F33615-77-C-1068 and by the U.S. Air Force Rome Air Development Center under Contract No. F19628-80-C-0150.

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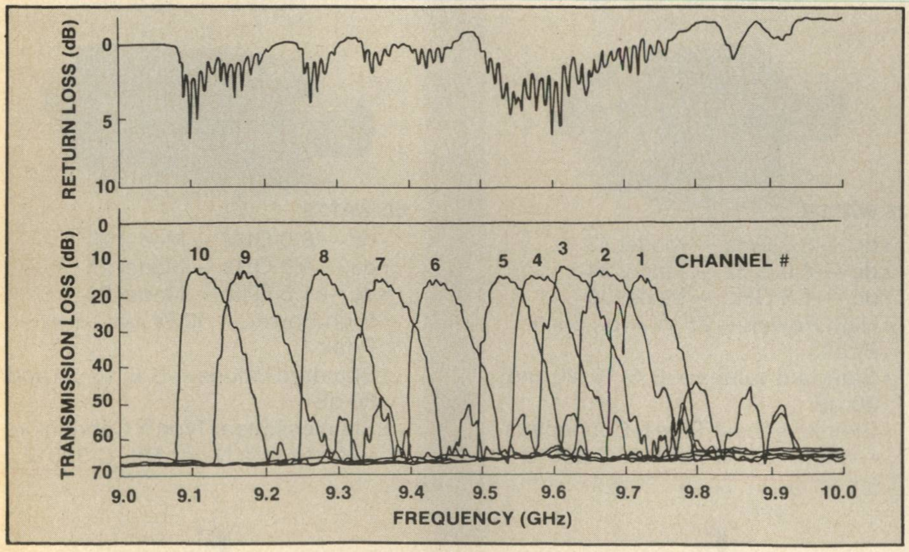

Fig. 10 Measured transmission loss and return loss from the input for 10-channel filter.

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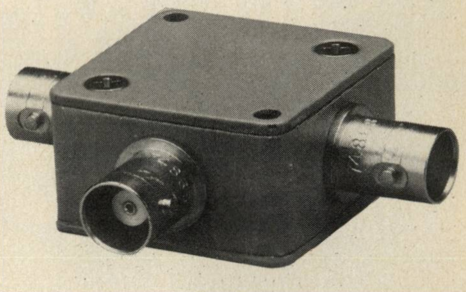
19.5 dB



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FREQUENCY (MHz) 0.1-2000		
COUPLING, db 19.5		
INSERTION LOSS, dB	TYP.	MAX.
one octave band edge	0.8	1.4
total range	1.5	2.3
DIRECTIVITY dB	TYP.	MIN.
low range	30	20
mid range	27	20
upper range	22	10
IMPEDANCE	50 ohms	

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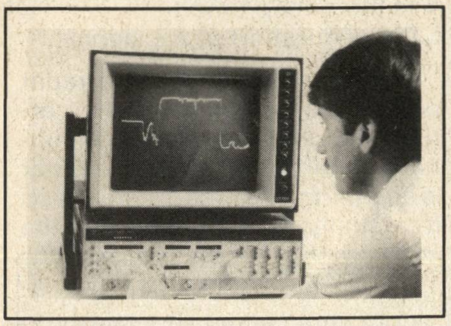
New Time-Domain Measurement Capability With Waveform Recorder/ Spectrum Analyzer Team

Douglas Nichols
Hewlett-Packard Company

A powerful measurement capability can be realized by combining a spectrum analyzer and an analog-to-digital waveform recorder. The spectrum analyzer's versality in receiving and down-converting incident microwave signals matches ideally with the waveform recorders ability to digitize and store an input. The set-up for the spectrum analyzer-waveform recorder combination is shown in Figure 1.

The primary benefit of adding a waveform recorder to accept video output from the analyzer is that the time-domain envelope of the rf input can be stored in the waveform recorder and processed, if desired.

Instead of using the spectrum analyzer in its most familiar mode of operation, where a filter is repeatedly swept across a selected frequency span (thus displaying time-limited frequency versus amplitude information), the spectrum analyzer is used in "zero span" mode. As this term implies, the filter is not swept in zero span mode. Instead, the filter remains stationary at a selected center frequency, with filter bandwidth controlled by the selected resolution bandwidth as shown in Figure 2. Thus, in zero span mode, the spectrum analyzer becomes a variable bandwidth super heterodyne tuneable receiver (the output is frequency-limited time versus amplitude information). The video output from a spectrum analyzer in zero span mode is the time-domain envelope of the rf input. It is this signal that is captured in the waveform recorder.



Hewlett-Packard's new HP 5108A Waveform Recorder, shown with the HP 1311A. Display samples input signals at rates to 20 MHz and stores resulting 10 bit codes in up to 16K words of memory.

The major benefits that the waveform recorder brings to the spectrum analyzer-waveform recorder combination are:

- The time domain envelope of the rf input can be captured, allowing analysis of the incoming microwave signals.
- Intermittent or transient phenomena, which wouldn't be seen on the spectrum analyzer, are captured by the waveform recorder.
- Pre-trigger data capture capability ensures that even a single rf burst will be completely sampled, not just the portion following the trigger.

A version of the spectrum analyzer-waveform recorder combination was implemented using an HP 8565A Spectrum Analyzer and an HP 5180A Waveform Recorder. All measurements shown in this article were made using these instruments.

Applications for this measurement capability are found in EW, ECM, radar, and noise measurement systems. Aside from these specific areas, true RMS measurements can be made using this set-up. These applications are detailed in following sections.

Probability Of Intercept

In all of the military-related applications for a spectrum analyzer and an analog-to-digital waveform recorder, the reason for using analyzer in the zero-span (time-domain) mode rather than the swept (frequency domain) mode centers around the concept of "probability-of-intercept".

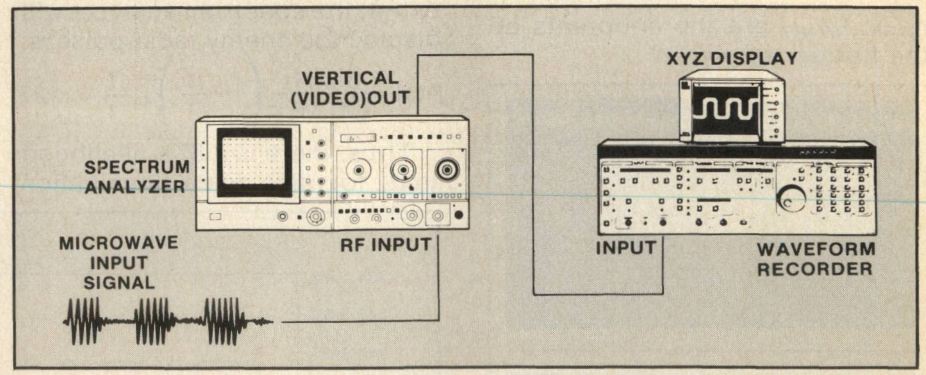


Fig. 1 Interconnections for spectrum analyzer-waveform recorder combination.

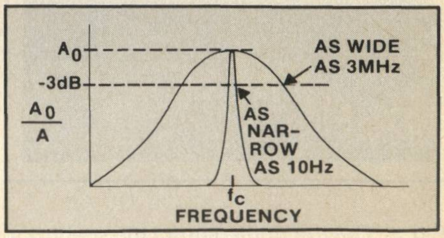


Fig. 2 An illustration of zero span mode and the wide range of passband filters which can be used with a fixed center frequency.

In its most general sense, probability of intercept is a measure of the likelihood of detecting a given signal with a given measurement system. For example, the probability of intercept of detecting a 35 GHz signal using a 1 GHz frequency counter is zero. Even if we use a 40 GHz counter the POI will probably not be 1, since the signal may be below the sensitivity of the counter. The goal, of course, is to use systems having POI's approaching 1.

When the spectrum analyzer is sweeping, the POI may be low, since the filter can only periodically pass through a given region of the spectrum. Quantitatively, the probability of intercept is given by

$$POI = \frac{f_{BW}}{f_{STOP} - f_{START}}\left(1 - \frac{t_S}{T}\right) + \frac{t_S}{T}$$

for $0 \leq t_S \leq T$ = 1 for $t_S > T$

where f_{BW} is the resolution bandwidth of the sweeping filter

f_{STOP}, f_{START} are the endpoints of the frequency sweep

T is the sweep time

t_S is the duration of the signal

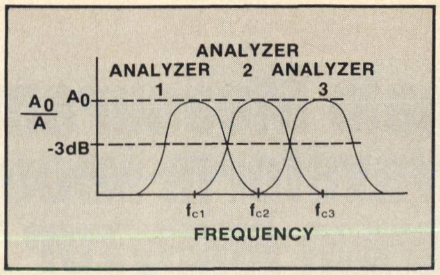


Fig. 3 Several analyzers used together in zero span mode to constantly monitor a wide frequency range.

For example, suppose a radar pulse train having a 10 GHz carrier frequency is used to illuminate an airplane. Since the radar is scanning, the time during which it actually is detectable by the plane will be short, say 100 ms. A spectrum analyzer aboard the plane is being used to constantly monitor the spectrum. If this analyzer is sweeping from 9.95 - 10.05 GHz in 1 second, with a resolution bandwidth of 3 MHz, the probability that the spectrum analyzer will display the enemy radar pulse is:

$$POI = \frac{3\ MHz}{100\ MHz}\left(1 - \frac{.1}{1}\right) + \frac{.1}{1} = .127$$

Thus, there is a 87% likelihood that, for any given illumination the signal will not be detected. The zero-span mode is an obvious way to increase POI. Although more hardware is needed, a group of spectrum analyzers could be placed in zero-span, with each analyzer centered at a frequency offset from the previous analyzer by the amount of filter resolution bandwidth. In this way, a section of spectrum can be covered with POI = 1, see Figure 3.

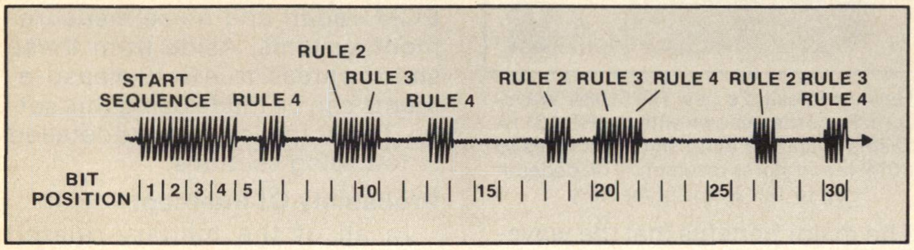


Fig. 4 The first 30 bits of coded microwave pulse pattern.

Pulse Pattern Recognition

There are many military (and non-military) applications where pulsed RF burst streams are generated in a coded pattern. The spectrum analyzer-waveform recorder is well suited for testing the code generating sources, or for comparing a received pulse pattern with a reference pattern (or even a library of reference patterns in order to identify it).

The spectrum analyzer is placed in zero span mode, the frequency is tuned to the pulse carrier frequency, and the resolution bandwidth is set as large as possible (to preserve the pulse shape as faithfully as possible). The down-converted rf pulse pattern is envelope (AM) detected and routed to the waveform recorder, where the signal is digitized and stored for processing.

To demonstrate the capability of the spectrum analyzer-waveform recorder combination for this application, a 5 MHz PRF, 2.819 GHz pulse carrier frequency bit stream was generated using a word generator to drive the PIN diode controlling the RF pulse

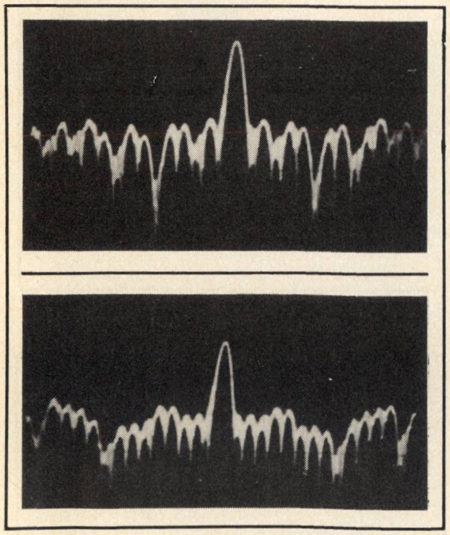


Fig. 5 The top photo shows the spectrum centered at 2.819 GHz for the example microwave pulse pattern. The lower photo shows the same pulse pattern with rule 2 (a pulse in every bit position) left out.

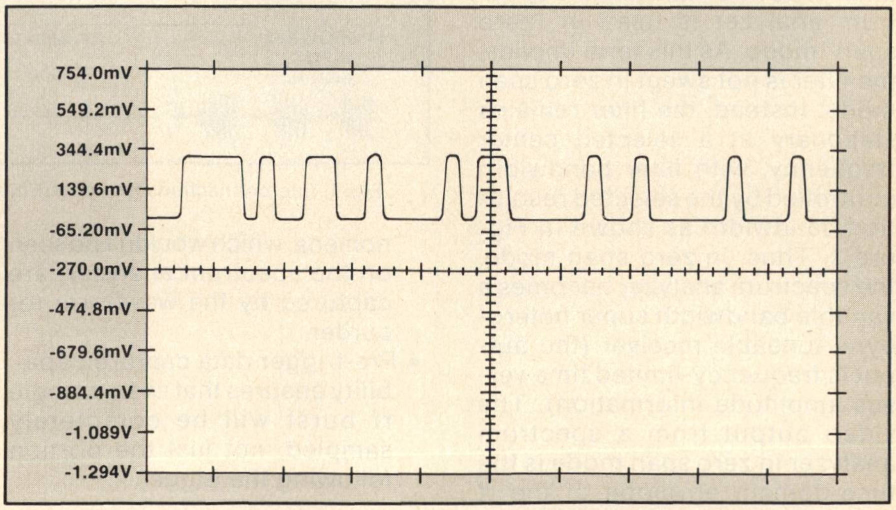


Fig. 6 Data automatically plotted after capture by HP 5180A Waveform Recorder. Data is the example coded microwave pulse pattern.

modulation. The beginning of the 256 bit pattern is indicated by four consecutive pulses following the begin sequence:

- A pulse is generated in every 9th bit position
- A pulse is generated in every 10th bit position
- A pulse is generated in the 6th bit position, 6 + 7th bit position, 6 + 7 + 8th bit position, etc..

The beginning of the sequence is shown in Figure 4.

The challenge here is to determine if a given pulse sequence matches this form. Trying to do the analysis in the frequency domain is unwieldy. Examples of the spectral content of the pattern and pattern without rule 2 (see Figure 5) show that, while it is clear that they are different, it is impossible to quantify *how* they are different.

By placing the spectrum analyzer in zero-span mode, centering the tuning frequency to 2.819

```
0: 512→N
1: dim X[0:N-1],Y[0:N-1]
2: ent "file of good pattern",r1
3: ent "file of test pattern",r2
4: ldf r1,X[*];ldf r2,Y[*]
5: N→D
6: for J=0 to N-1
7: for I=0 to N-1
8: if abs(X[I]-Y[(I+J)modN])>25;C+1→C
9: next I
10: dsp "SHIFT=",J,"CORR.=",(N-C)/N
11: if C<D;C→D;J→E
12: if C=0;gto 15
13: 0→C
14: next J
15: scl 0,N-1,-1000,3800
16: plt 150,3720,1;csiz 3;lbl "Waveform Pattern Analysis"
17: plt 10,3550,1;csiz 2;lbl "reference waveform"
18: for I=0 to N-1
19: plt I,X[I]+2500
20: next I
21: pen;plt 10,2350,1;lbl "actual best fit waveform"
22: for I=0 to N-1
23: plt I,Y[(I+E)modN]+1300
24: next I
25: pen;plt 10,1200,1;lbl "difference waveform"
26: for I=0 to N-1
27: plt I,Y[(I+E)modN]-X[I]+130
28: next I
29: plt 10,-950,1;lbl "Total differences for ",N,"samples= ",D
30: end
*6712
```

DATA ENTRY

THIS SECTION CONTINUALLY ADJUSTS PHASE OF TWO WAVEFORMS UNTIL BEST FIT IS DETERMINED.

PLOTTING ROUTINES

HP 9825 controller listing for Fig. 7 patterns.

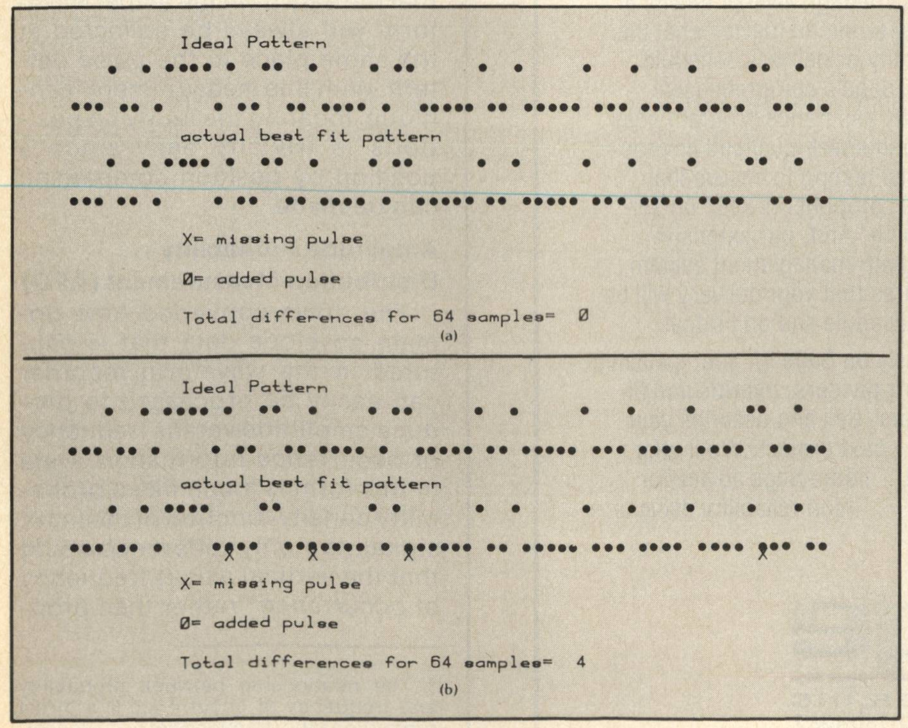
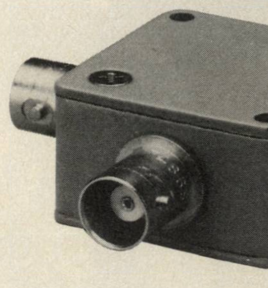


Fig. 7 In 7(a) the captured pattern matched the reference, hence, no errors detected. In 7(b) rule 4 was deleted, hence the missing pulses.

[*Continued on page 106*]

10.5dB directional couplers



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total range	1.0	1.3
DIRECTIVITY dB	TYP.	MIN.
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mid range	33	25
upper range	22	15
IMPEDANCE	50 ohms	

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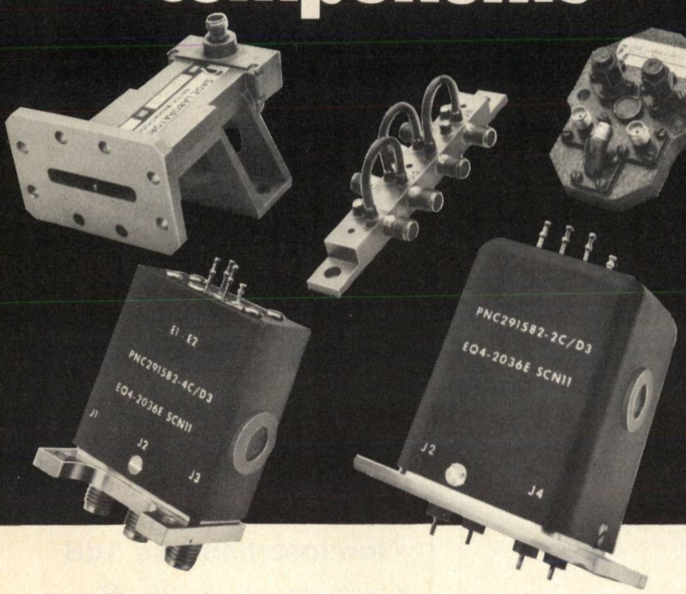

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[From page 105] **TIME-DOMAIN**

GHz, and using maximum filter resolution bandwidth (3 MHz), the time domain pulse stream envelope was captured in the waveform recorder. In this case, the waveform recorder sample rate was controlled externally, sampling once per bit position. This is the most efficient way to sample the data because only one sample is needed to determine the status of each bit position. Using the HP 5180A Waveform Recorder, this means that a sequence of up to 16,384 pulses could be stored in one measurement. A less efficient, but still valid way to collect the data is to use an internally set sample rate, ensuring that this rate exceeds the pulse rate. Examples of data collected using both of these sampling methods is shown in Figure 6.

By adding a controller to the set-up, the data collected in the 5108A can be compared to the intended (or ideal) waveform. Figure 7 shows output for several waveforms that were actually measured and compared with the ideal waveform.

An analysis program shifts the relative phase of the reference and actual waveforms to find the best possible fit. This is necessary since there is no way to ensure that the data from the actual waveform will always be collected at the same place in the pulse pattern. With this measurement technique, locating the faulty bit positions is trivially easy since a position by position comparison can be made.

Amplitude Probability Distribution Measurement (APD)

The down-converted time domain envelope data that is captured in the waveform recorder can easily be processed to produce amplitude versus frequency of occurrence information. Data in this form is much like a probability density function of the input signal, the only difference being that the vertical axis is frequency of occurrence* rather than prob-

* The relationship between probability and frequency of occurrence is a direct proportionality: (frequency of occurrence)/(total samples) = (Probability of occurrence)

ability of occurrence. Figure 8 shows a simple example of the results of transforming a sinusoidally AM modulated microwave signal into an amplitude probability distribution.

Using APD In Surveillance

Surveillance is the business of monitoring signals of unknown composition and extracting whatever information these signals contain. This information can take many forms and APD can be used to assist in the identification process.

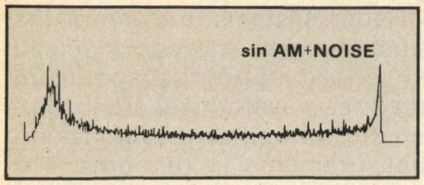

Fig. 8 Amplitude probability distribution (APD) of a microwave signal amplitude modulated by a sine wave.

Suppose that a signal of unknown composition has been detected and is suspected of being a voice transmission. Before the signal can be demodulated, it is necessary to determine what type of modulation was used to generate the transmission. APD analysis can be used to identify the modulation type, since each modulation scheme will exhibit a characteristic amplitude probability distribution. The technique is especially useful when many signals are present simultaneously, because the APD analysis can sort out the various signals, whereas the frequency or time domain signal may be too complex for analysis. Figure 9 shows the type of results that can be obtained on multiple signal inputs.

Using APD For Noise Measurements

One of the most difficult problems in RFI (Radio Frequency Interference) and EMI (Electro-Magnetic Interference) measurement is to find a detector whose response does not mask the process that is being measured. For instance, a quasi-peak detector (commonly found in EMI measuring instruments), provides output that is a weighted average most sensitive to the peak voltage and small levels below the peak. Such a detector cannot be used to measure the rapidly changing amplitudes encountered with noise. What is needed in noise measurement is a true RMS detector. The spectrum analyzer-waveform recorder can provide this detection capability.

The data for the noise process being monitored is sampled, perhaps many records of waveform recorder data are collected, and an APD is generated. In EMI and RFI noise analysis, the APD of the noise is the end goal, since it is such a useful characterization of the noise process. Since these measurements are always specified at some frequency with a particular bandwidth, the spectrum analyzers variable resolution bandwidths are used to control the input and assure that the waveform recorder is capturing the proper signals.

If the noise process being monitored is broadband, a wide dynamic range measurement is needed, since broadband noise is usually characterized by a con-

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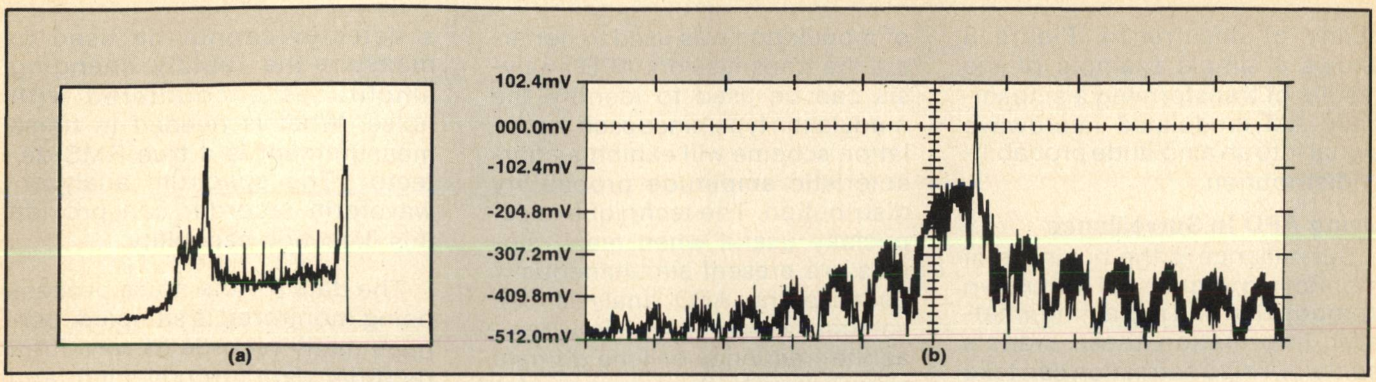


Fig. 9 A sinusoidally modulated microwave signal and another CW microwave signal generated the APD of 9(a). 9(b) shows the corresponding spectrum which was digitized by the 5180A and plotted on a 9872C plotter.

tinuous low level occasionally punctuated by strong noise bursts. For example, atmospheric noise typically exhibits amplitude variations as great as 90 dB. Since spectrum analyzers can provide displayed dynamic ranges of 90 dB the spectrum analyzer-waveform recorder combination is ideal for broadband noise monitoring.

Use The Quantized Data To Compute True RMS Values

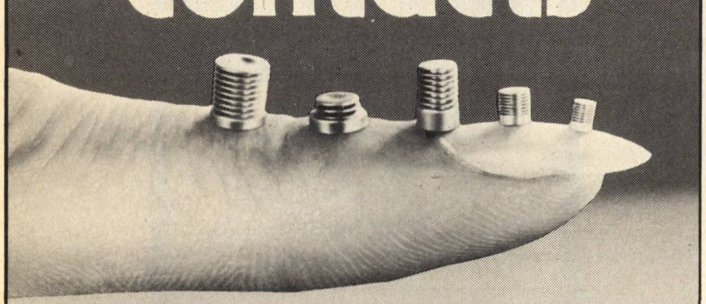
As was mentioned in the previous section, the spectrum analyzer-waveform recorder combination can provide true RMS detection capability. This is so because the waveform recorder captures a point-by-point copy of the input signal. After outputting the data to a controller, the samples may be converted to power, a mean value computed, and the square root taken. The result is a *true* RMS measurement. Such measurements are especially useful on signals having unknown statistical properties, since filters that can be made to correct for known statistical characteristics are not applicable to the signals.

Radar

Radar systems are one of the most common sources of pulsed RF energy and the spectrum analyzer-waveform recorder combination is well suited for making measurements in this area. For pulse measurements, the analyzers maximum resolution bandwidth should be chosen to allow best possible reconstruction of the original pulse. Using an analyzer with a 3 MHz maximum resolution bandwidth allows pulses having widths down to 500 nanoseconds to be faithfully repro-

Electrodeposited bellows

contacts



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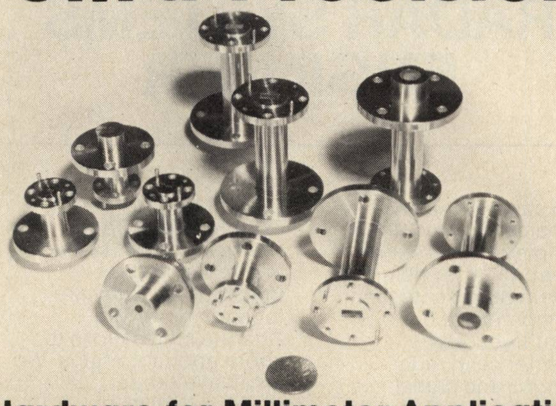

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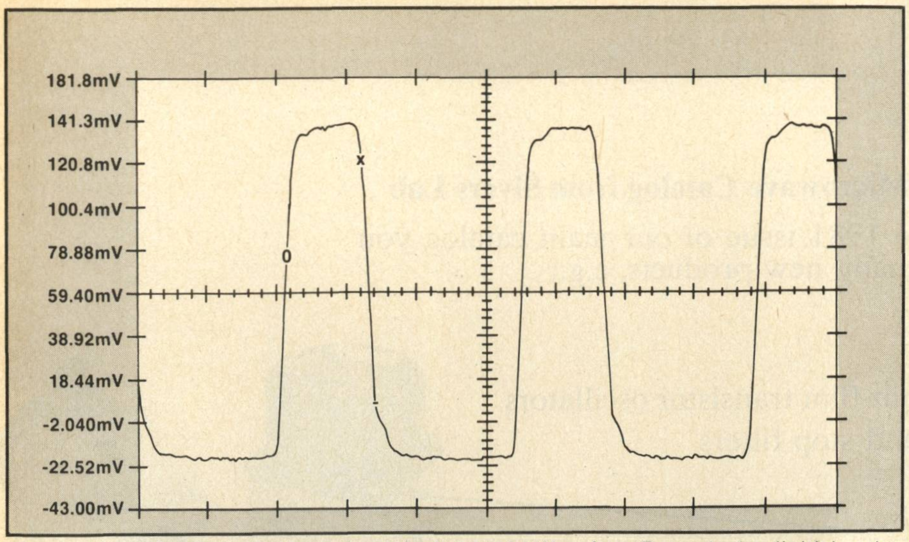

Fig. 10 Microwave pulse train captured by the 5180A Waveform Recorder by digitizing the video output of a spectrum analyzer used in the zero span mode.

duced. This value of pulse width is a good minimum for the waveform recorder as well, since a 50 nanosecond sample rate would obtain 10 samples during the pulse which is a practical minimum for good visual reconstruction of the demodulated pulse. Figure 10 shows a pulse train captured using an HP 8565A Spectrum Analyzer and an HP 5180A Waveform Recorder. A major advantage of using the waveform recorder to capture the pulses is that, once armed, the waveform recorder will capture any signal that appears. A single pulse burst can be captured as well as a repeating pulse train. Thus signals that may suddenly appear for short durations will not be lost by the waveform recorder. Another measurement that is useful on Moving Target Indication (MTI) radar systems is measurement of the pulse to pulse amplitude jitter, a term in the MTI improvement factor. The improvement factor in an indication of a particular MTI system's capability to detect a moving target return from a high level of stationary clutter (noise) return and has a 20 log ($A/\Delta A$) effect on this factor, where A is the pulse amplitude and ΔA is the pulse to pulse (interpulse) amplitude change. Typical requirements for amplitude jitter call for jitter no worse than 40 to 50 dB. The HP 5180A Waveform Recorder can resolve amplitude to better than 60 dB (.1%), so it can make the amplitude jitter measurement.

The applications cited do not begin to exhaust the applications of the waveform recorder-spectrum analyzer combination. The frequency and time domain acquisition capabilities that are brought together by these two instruments provide solutions to a wide range of measurement problems. ■

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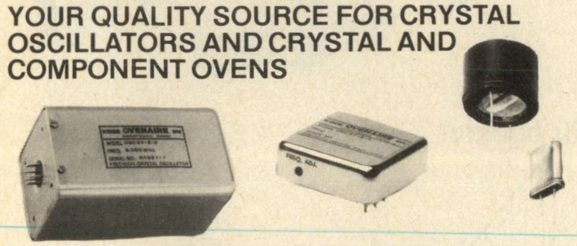

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[From page 38] MSW

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Typical S-parameters in 50 ohm systems for packaged types are shown in Table I.

Typical performance is shown in Table II.

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Circle 172.

TABLE I
TYPICAL S-PARAMETERS IN 50Ω SYSTEMS FOR PACKAGE TYPES

FREQUENCY (MHz)	S21 [MAGNITUDE/ANGLE (°)]					
	MA-4F001-297		MA-4F001-297 Source Grounded		MA-4F100	
2000	2.41	122.4	3.31	101.1	2.72	76.0
3000	2.18	96.7	2.56	73.6	2.14	40.0
4000	1.92	73.1	2.03	50.5	1.68	8.0
5000	1.70	51.4	1.66	30.1	1.48	-25.0
6000	1.52	31.2	1.39	11.3	1.40	-61.0
7000	1.37	12.0	1.19	-6.3	1.30	-95.0
8000	1.24	-6.5	1.03	-23.0	1.20	-133.0
9000	1.13	-24.5	.90	-39.0	1.05	-174.0
10000	1.04	-42.1	.79	-54.4	—	—
11000	.97	-59.7	.71	-69.2	—	—
12000	89	-77.2	.63	-83.4	—	—

TABLE II

	Part No.		
	MA 4F-001	MA 4F-002	MA 4F-100
Freq. (GHz)	10	10	6
G_1 (dB)	7	6	6
P_{01} (dBm)	17	23	23
J_{MAX} (°C/W)	150	60	40
Pkg. Style	-297	-297	-259
Price (10-99)			
Chip	$15	$30	$30
Packaged	$30	$45	$45

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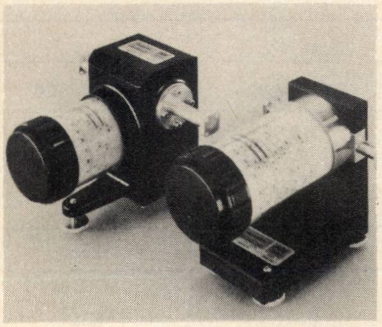
Design improvements have been completed on the Hitachi precision direct reading frequency meters and attenuators manufactured and sold by Hughes Aircraft Company, and those instrumentation products are now in full production at the Hughes Electron Dynamics Division. The instruments, available from stock in eight waveguide bands between 26.5 aznd 166 GHz, display improved accuracy and durability as a result of the Hughes productization program.

The accuracy of the frequency meter measurements has been improved (see Table I) by increasing the dip depth while maintaining a high Q. The frequency meter cavities are now precisely dimensioned silver plated electroformed parts assuring the highest practical Q. The cavities have been characterized and computer modeled to minimize errors due to cavity anomalies. The new electroformed cavities and computer modeling supplied a unique opportunity: The error at key frequencies (35 GHz in Ka-band, 60 GHz in V-band, 94 GHz in W-band, and 140 GHz in D-band) can be made as low as 0.05 %. Careful characterization of the cavities also yields the deepest practical "dip". Figure 1 compares the dip characteristics of the new and old versions.



Ka frequency meter and attenuator.

The accuracy of the attenuator has been improved by taking advantage of numerically controlled machining techniques to assure repeatability of critical part dimensions and by ruggedizing the designs.

Durability of both the attenuator and frequency meter has been improved by strengthening the waveguide flanges and simplifying calibration and maintenance procedures. Since these instruments in a typical test set up are connected to components not self supporting, stresses at the flanges occur causing bending or breakage. The strength has been increased by encapsulating the waveguide in brass sleeves with integral flanges. The durability of the finish has also been improved by using an anodizing process.

The Hughes 4571xH series of precision direct reading frequency meters are priced from $1265 to $1870 in Ka through W-bands (26.5 to 110 GHz), at $2970 in F-band (90-140 GHz) and at $3960 in D-band (110-170 GHz). The Hughes 4572xH series of precision direct reading attenuators is priced at $1450 in Ka-band (26-5-40 GHz) through E-band (60-90 GHz), at $1485 in W-band (75-110 GHz), at $1870 in F-band, and at $1980 in D-band. Delivery of both products in all bands is from stock. ■

Circle 131.

TABLE I

FREQUENCY METER ACCURACY SPECIFICATIONS

BAND	OLD SPEC (% max.)	NEW SPEC (% max.)	NEW TYPICAL (%)
Ka	0.14	0.12	
Q	0.23	0.12	
U	0.25	0.15	
V	0.33	0.20	
E	0.34	0.20	0.10
W	0.47	0.25	
F	0.50	0.50	
D	0.50	0.50	

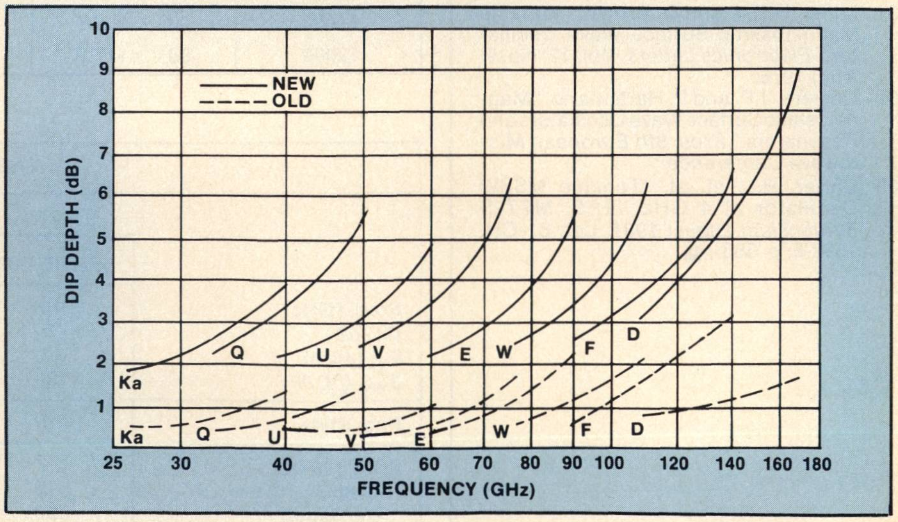


Fig. 1 Dip-depth vs. frequency of old and new frequency meters.