$S/N + D$ and thermal noise caused in the noise loading test of 1380 channels were more than 75 dB and more than 78 dB, respectively.

CONCLUSION

A 4-GHz all solid-state 1380-channel FM transmitter using 5 W in total dc input, 220 mW in output power, and exhibiting more than 4 percent in dc-to-RF signal-conversion efficiency has been developed by using, in combination, a low-power local oscillator, a low-level up converter, and a high-gain transistor injection-locked amplifier.

Using only two transistors and a varactor diode together with a stabilizer and an alarm circuit, this combination provides a reliable and practical FM transmitter.

ACKNOWLEDGMENT

The authors wish to thank Dr. Y. Kaito for his valuable discussions and suggestions on this work. They also wish to thank I. Haga for his preparation of the transistor injection-locked amplifier and local oscillator.

REFERENCES

[1] a) P. T. Chen, "Design and applications of 2 ~ 6.5 GHz transistor amplifiers," in *1973 IEEE Int. Solid-State Circuits Conf. Digest Tech. Papers*, pp. P76–P77.
b) A. Presser *et al.*, "2.6 ~ 3.2 GHz transistor amplifier," in *Proc. 1971 High Frequency General Conf.* (Cornell Univ., Ithaca, N. Y., 1971), vol. 3, p. 445.

Development of a Pulse Compression Distance Measuring Equipment System Using Surface Acoustic Wave Devices

DONALD W. MELLON AND WILLIAM D. DANIELS, MEMBER, IEEE

***Abstract*—Proposed *C*-band distance measuring equipment (DME) requires the use of triodes to generate high-power *C*-band transmitted pulses. The inherent short life of these triodes necessitates placing this equipment in physically accessible areas of the aircraft, often long distances from the antenna. Losses incurred in transmitting the *C*-band pulse to the antenna can increase the power requirements of the system.**

A pulse compression system has been designed to solve the cost of maintenance problems by using reliable low-power solid-state transmitters and also allowing the equipment to be installed close to the antenna. The use of a pulse compression system allows a reduction in peak transmit power by a factor equal to the time bandwidth (BW) product of the transmit pulse. Size, weight, and reliability are also improved by implementing surface wave devices (SWD's) in the pulse compression airborne interrogator.

INTRODUCTION

Distance measuring equipment (DME) systems aid aircraft by providing long-range navigational data and short-range high-resolution landing information. Current systems, operating at *L*-band frequencies, do not meet the accuracy requirements of the proposed microwave landing system (MLS) [1]. A pulsed RF *C*-band system, identical in all significant respects to the *L*-band system except for channel bandwidth (BW), has been designed to meet these range resolution requirements.

A significant undesirable feature of the *C*-band system is the air to ground power requirements which necessitates the use of microwave triodes which typically have mean time between failures around 4000 h. Thus the tubes must be placed in easily accessible areas of the aircraft for maintenance. Available placement areas are long distances from the antenna, thus increasing the power generating requirement.

In this short paper we discuss an alternate *C*-band DME system which reduces the transmitted power requirements by a factor of thirty (30), thus allowing the use of low-power high-reliability [2] solid-state circuitry while providing the accuracy requirements of the MLS system. The pulse compression airborne equipment will occupy one half the volume and weigh one third as much as the pulsed *C*-band equipment. Due to reduced periodic maintenance this equipment can be located based on performance rather than accessibility.

SYSTEM DESCRIPTION

The basic elements of a DME system are the airborne interrogator, ground based transponder, and an airborne clock. Calculating the distance between interrogator and transponder is accomplished by determining the signal propagation time for the round trip. A block diagram of the complete surface wave device (SWD) DME system is shown in Fig. 1.

Interrogation of a particular ground station is done by transmitting a pulse pair and timing the reply interval. Reply identification is done in the aircraft by using random pulse position jitter and using this *a priori* knowledge for range gating the reply. That is, the aircraft transmits fixed pulse pairs at a random repetition rate and then processes only those replies which occur in a predicted range gate. Ground station channelization is done in the time domain by 10 different pulse pair spacings and in the frequency domain by 20 separate frequency bands. A description of present DME systems is given by Hirsch [3].

Although the performances of pulsed *C*-band DME and the pulse compression DME are similar, there are important implementation differences. First, the peak power requirement of the triode DME is 150 W to provide 30-nm range whereas the pulse compression system using SWD's with a 3-MHz BW and 10-μs expanded pulse length requires only 5 W of transmitted power. This power can easily be generated using a 40-W solid-state *L*-band amplifier and X-4 multiplier. Second, the signal dynamic range requirements dictate the use of a precision 60-dB dynamic range log amp in the pulsed DME receiver, whereas in the SWD system all signals are processed through a dynamic range equal to the signal to noise improvement ratio of 13 dB. Third, the spectrum BW of the transmitted RF pulse is several times larger than the channel BW requiring a frequency validation circuit in the receiver which limits adjacent channel selectivity to 8 dB. The SWD spectrum BW is concentrated resulting in an adjacent channel selectivity of 30 dB. Fig. 2 is a block diagram of the SWD DME interrogator and transponder IF sections.

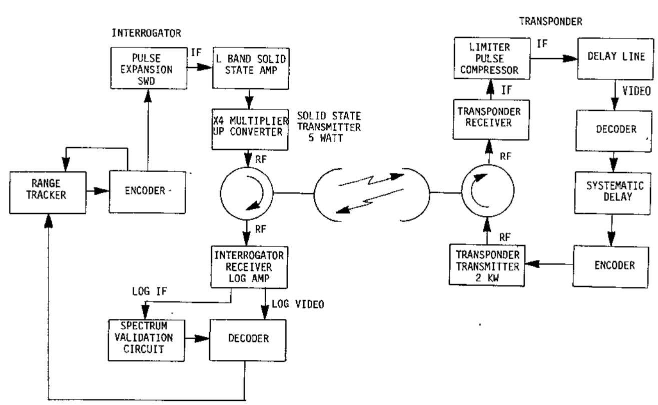

Fig. 1. Block diagram of the complete SWD DME system.

Manuscript received May 21, 1974; revised August 30, 1974.
The authors are with Texas Instruments, Incorporated, Dallas, Tex. 75222.

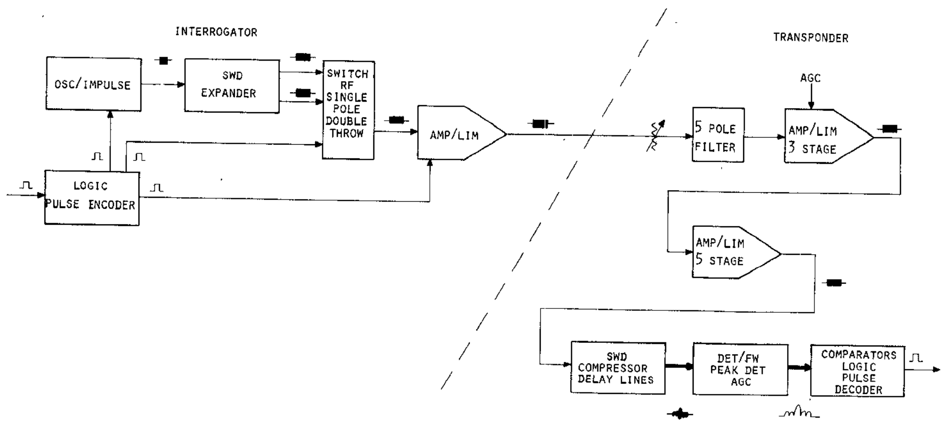

Fig. 2. Block diagram of the SWD DME IF sections.

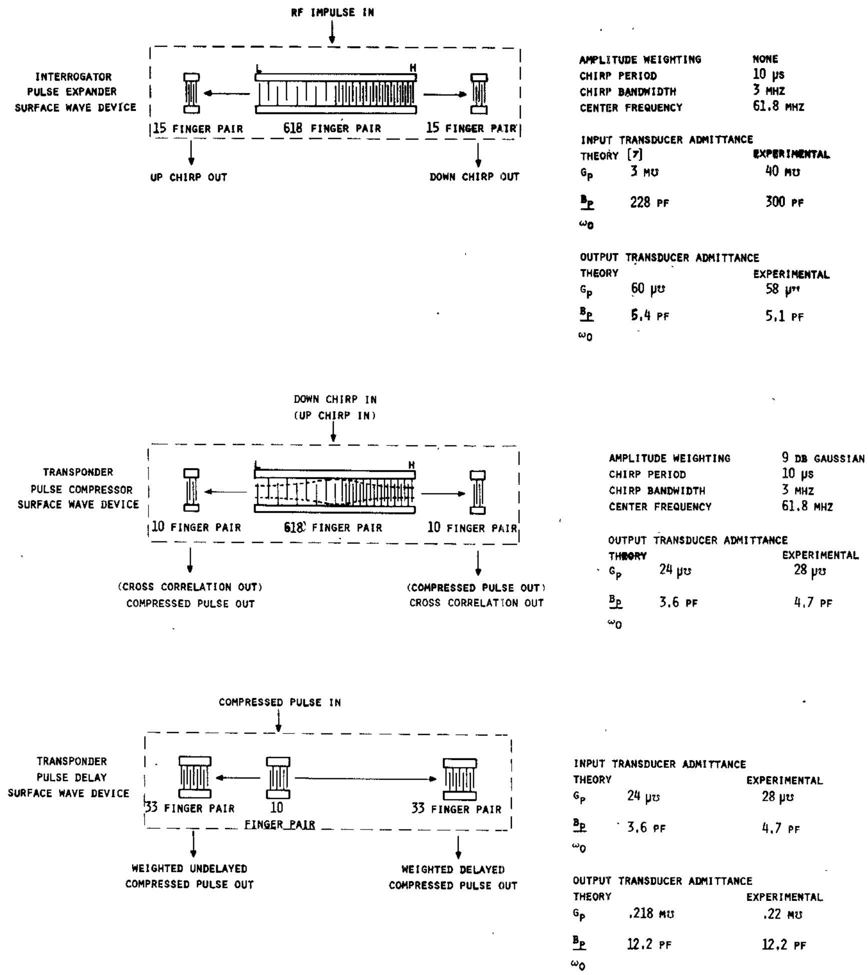

Fig. 3. SWD's used.

SWD DESIGN

The pulse expansion/compression system used linear FM filters designed to process signals 10 μs in length with a BW of 3 MHz centered at 61.8 MHz. To reduce surface wave distortion due to mechanical reflections a split finger transducer design [4] was chosen and fabricated on 42-3/4° y-rotated x-propagating quartz substrate (ST cut). The substrate material is a well-defined inexpensive readily available temperature stable material [5]. Fig. 3 shows the geometry of all of the SWD's and lists the theoretical and experimental center frequency admittances.

The surface wave compressor has amplitude weighting (apodization) to reduce the time sidelobes to -26 dB at the system level. The sidelobe reduction is obtained by using a 9-dB Gaussian weighting on a 10-μs dispersive transducer in series with a nondispersive narrowband transducer. This design provides the best combination of sidelobe reduction and transducer impedance levels to meet system performance while also reducing matching network complexity.

Fig. 4 compares the theoretical and experimental linear FM spectrum. Fig. 5 shows the theoretical and the experimental compressed pulse and cross correlation pulse. Table I compares the theoretical and measured design parameters.

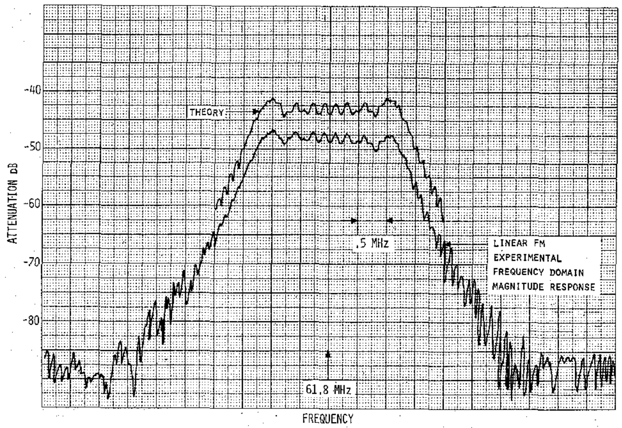

Fig. 4. Theoretical and experimental spectrum.

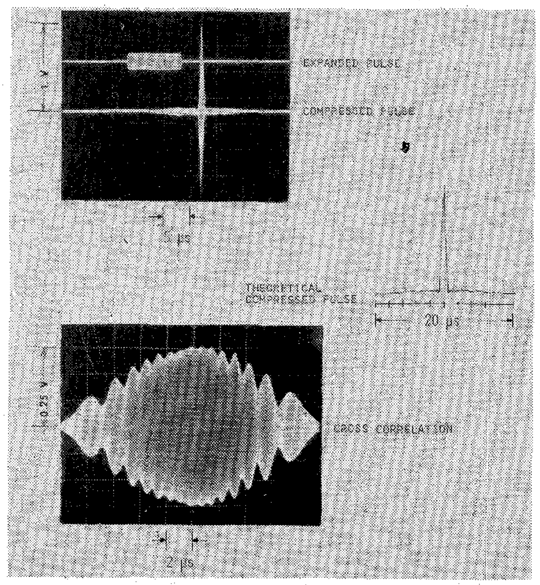

Fig. 5. Theoretical and experimental compressed pulse and cross correlation pulse.

TABLE I
DESIRED SWD PARAMETERS AND EXPERIMENTAL RESULTS

PARAMETER	DESIRED	OBTAINED
CENTER FREQUENCY (MHz)	61.8	61.8
PULSE LENGTH (μs)	10.0	10.0
BANDWIDTH (MHz)	3.0	3.0
CW INSERTION LOSS UNMATCHED (dB)	41	47
SIDELOBE LEVEL (dB)	-30	-28
PULSE WIDTH, 3 dB (ns)	400	410
PHASE ERROR (ppm)	0	110
CROSS CORRELATION LEVEL (dB)	-12.8	-12.3
CROSS TALK, BELOW SIDELOBES (dB)	10	15

To achieve the performance of Table I it was necessary to compensate for a major SWD distortion effect similar to triple transit. Triple transit is caused by acoustic power being reflected between input and output transducers and occurs because SWD transducers are three port devices which cannot be simultaneously impedance matched at the electrical and acoustic ports, therefore reflecting incident energy at the unmatched ports. If the electrical port is matched, then optimum power transfer is obtained at the expense of a large reflected acoustic wave. This problem becomes more critical if multiple output transducers are used because reflected energy can reach output transducers after suffering only a single reflection rather than being attenuated through two reflections as in triple transit. This is depicted in Fig. 6. For this reason it is necessary to determine the matching network for each electrical port which will result in allowable reflected energy levels [6].

The problem restated is: given an acceptable level of reflected acoustic energy (S_{11}) from a given port, determine the electrical source or load admittance of that port which will allow optimum power transfer (S_{13}). For the SWD input the desired value of G_s is obtained by maximizing $|S_{13}|$ with respect to S_{11}, yielding

$$\frac{G_s}{G_a} = \left[\frac{1}{|S_{11}|^2} - \left(\frac{B_s + B_a}{G_a}\right)^2\right]^{1/2} - 1.0 \tag{1}$$

where G_a is the acoustic conductance, B_a is the acoustic susceptance, B_s is the source susceptance, and G_s is the source conductance. The reflected energy $|S_{11}|$ is plotted versus G_s/G_a for several values of $(B_s + B_a)/G_a$ in Fig. 7. For example, if a reflected wave is required to be 20 dB below the incident wave and a tuned match is desired ($|B_s + B_a| = 0$), a value of $G_s/G_a = 9.2$ is required to achieve that level while producing maximum S_{13}.

An additional distortion effect was a result of a slight asymmetry which exists between up chirp and down chirp pulse compression SWD's. Acoustic energy traveling under electrodes spaced corresponding to frequencies higher than the frequency of the acoustic disturbance is distorted by the electrodes less than if the spacing corresponds to lower frequencies. This causes an effective center frequency differential between up and down chirps. The chirp SWD's were analyzed on an equivalent circuit model computer program and were compared to experiment. This comparison of the modeled phase slope at the band edges predicts a 100-ns correlation peak difference on up and down chirps and indeed a difference of 100 ns was observed experimentally. This error is, however, averaged to zero in the same manner as other bias errors to be discussed later.

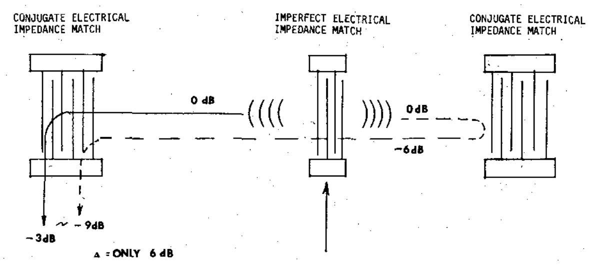

Fig. 6. Delay line reflections with multiple output transducers.

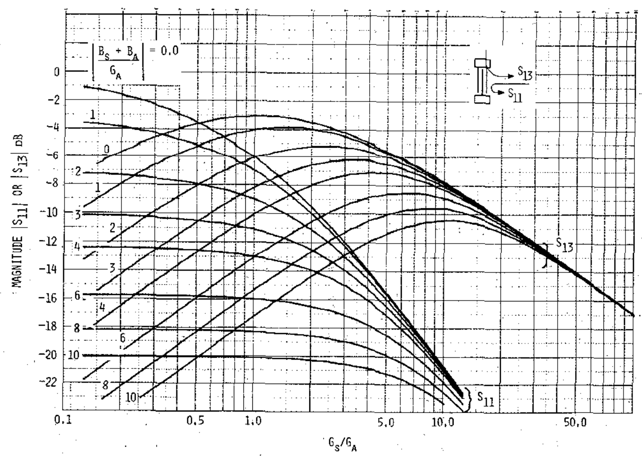

Fig. 7. Suppression of reflected acoustic wave $|S_{11}|$ and CW insertion loss per transducer $|S_{13}|$ as a function of source conductance G_S, with source susceptance B_S and acoustic susceptance B_A as parameters.

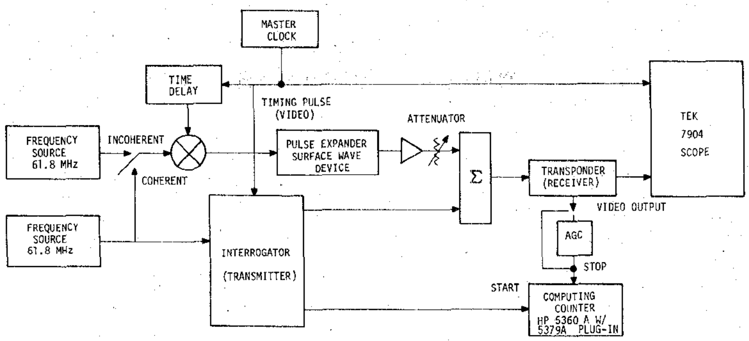

Fig. 8. Block diagram of equipment setup for experimental multipath effects.

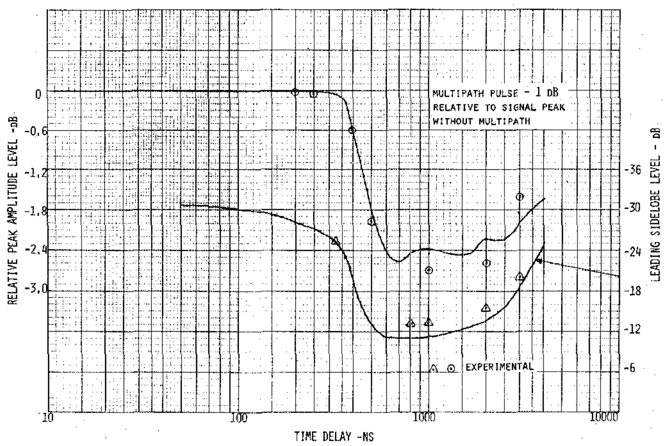

Fig. 9. Pulse suppression and leading sidelobe level versus multipath time lag.

DISCUSSION OF SYSTEM ERRORS

Only random errors need be considered for DME systems since steady-state bias errors can be calibrated out. For pulsed RF L-band and C-band systems the important errors result from multipath signals, jitter in thresholding, noise jitter, jitter in the clock-counter, and jitter due to processed replies from other interrogators [3]. The total distance error is a function of the rise time of the RF pulses, the receiver BW's, and the integration time (actually the number of returns over which the replies are averaged). At C band, faster rise times and wider BW's provide higher accuracies than L-band equipment. Therefore, C-band DME's may obtain the desired level of accuracy by averaging fewer replies and hence can handle a greater density of traffic.

Doppler shift, local oscillator drift, and temperature induced phase errors were eliminated in this system by alternately transmitting and detecting up and down chirp expanded pulses. The correlation peak position as a function of time for a linear FM compressed pulse is given by

$$t = t_0 \pm (T/\mathrm{BW})\ \Delta f \tag{2}$$

where t_0 is the theoretical or zero error position, T is the pulse length, BW is the bandwidth, and Δf is the mismatch between the center frequency of the received pulse and the center frequency of the SWD correlation filter caused by the bias errors mentioned earlier. Since for an up chirp the Δf caused by any of these effects will have the opposite sign from that of a down chirp, the average error for detection of both chirp directions is zero.

Multipath distortion effects, caused by processing reflected interrogation signals, are the worst single source of error for either pulsed RF or pulse compression DME systems. This error in pulse compression systems has been investigated both theoretically on a hybrid computer and experimentally using the breadboard DME IF section shown in Fig. 8. The measurements were made without translation to C band since the multipath signals will travel many more wavelengths than the true signal and the phase difference will vary from pulse to pulse as the aircraft moves. These incoherent signals were realized experimentally by generating chirps from separate expanders, summing in a limiter, and compressing to observe the multipath distortion. The compressed pulse resulting from two time overlapping chirp pulses is not merely a superposition of two signals because the hard limiter will remove any amplitude ripple.

Multipath effects in pulse compression systems cause time of arrival (TOA) errors by suppressing and time shifting the compressed pulse. In addition, large multipath signals can produce leading time sidelobes only 11 dB below the compressed pulse peak. (See Fig. 9.) The TOA errors caused by the suppression and distortion of the compressed pulse as well as false triggers from large leading sidelobes can be reduced by using a floating threshold system. However, TOA errors caused by time shifts of the compressed pulse cannot be easily reduced. For a given multipath time delay difference ΔT and amplitude difference ΔM the resulting TOA will depend on the phase difference between the signals which, as mentioned, is random and varies from pulse to pulse. If TOA measurements were averaged over many returns under these conditions a net positive or negative bias error would result. Fig. 10 shows the mean ($\bar{x}$) per pulse bias error and its standard deviation (σ) as a function of ΔT for several values of ΔM. Fig. 11 shows the bias error as a function of phase for several values of ΔT and ΔM. Averaging up and down chirps will not eliminate these errors.

The analysis where the interfering signal is only 1 dB below the desired signal represents a severe case. In actuality, the situation is dynamic and when multipath exists a typical value for $\boldsymbol{\Delta}m$ of 6 dB is expected.

CONCLUSIONS

A pulse compression DME system using dual SWD (up/down) chirp transmission in the airborne interrogator and SWD compressors and delay lines in the transponder can equal the accuracy of the proposed pulsed CW C-band system while reducing cost of maintenance and offering size and weight reductions. This was accomplished by reducing the power requirements by a factor of 30, allowing solid-state devices to replace the triode tubes in the transmitter.

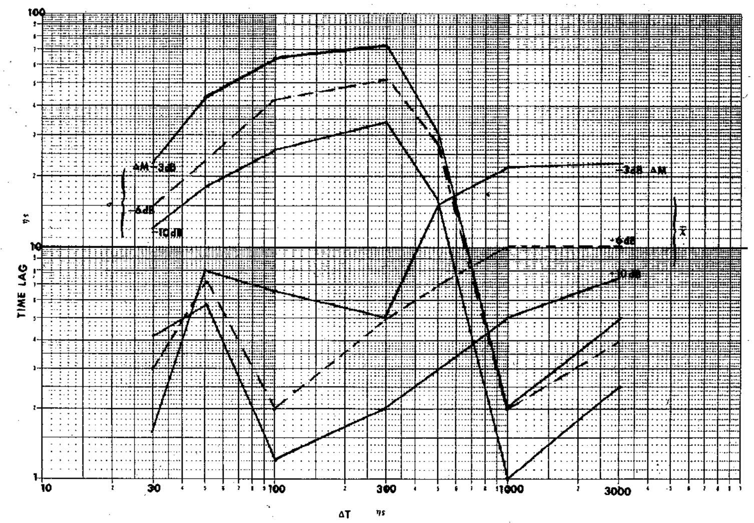

Fig. 10. Experimental mean ($\bar{x}$) and standard deviation (σ) per pulse bias error as a function of multipath time delay difference ΔT for several values of multipath amplitude difference ΔM.

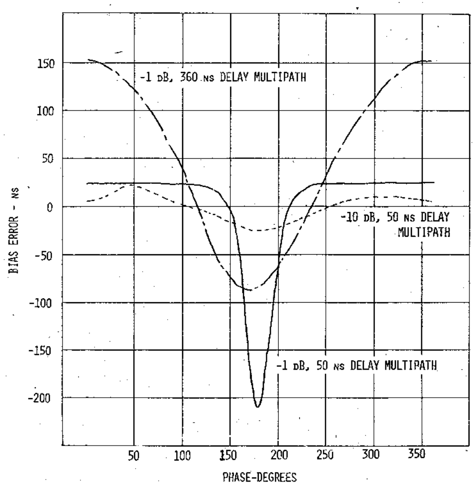

Fig. 11. Bias error versus multipath phase.

Additional benefits include a 30-dB adjacent channel selectivity rather than 8 dB and a requirement for a receiver with only 13 dB of dynamic range instead of 60 dB.

A detailed evaluation of system errors unique to pulse compression systems shows that a floating detection threshold system using alternate up and down chirp transmission will have no sources of error not present in the existing and proposed pulsed CW systems. TOA errors due to multipath have been experimentally determined and shown to be less than 24 ns (12 ft) for expected worst case conditions.

ACKNOWLEDGEMENT

The authors wish to thank J. Couvillon and B. Maher for the many technical contributions they made. The computer simulation was done by C. T. Cadenhead.

REFERENCES

[1] RTCA SC-117, "A new guidance system for approach and landing," Document DO-148, vol. 1, Dec. 18, 1970.
[2] R. Sudbury, "Operation of solid state transceivers in a *L*-band array," in *Proc. 1974 IEEE S-MTT Microwave Symp.* (Atlanta, Ga.), 1974, pp. 65–67.
[3] C. J. Hirsch, "*L*-band DME for the microwave-landing-system," Final Rep., Contract W1-3086-1, Feb. 1972.
[4] T. W. Bristol *et al.*, "Application of double electrodes in acoustic surface wave device design," in *Proc. 1972 IEEE Ultrasonics Symp.*, Oct. 4–7, 1972, Paper CHO 708-8 SU, pp. 343–346.
[5] D. T. Bell, "Phase errors in long surface wave devices," in *Proc. 1972 IEEE Ultrasonics Symp.*, Oct. 4–7, 1972, pp. 420–423.
[6] H. M. Gerrard, "Acoustic scattering parameters of the electrically loaded interdigital surface wave transducer," *IEEE Trans. Microwave Theory Tech.* (Corresp.), vol. MTT-17, pp. 1045–1046, Nov. 1969.
[7] C. S. Hartmann, D. T. Bell, Jr., and R. C. Rosenfeld, "Impulse model design of acoustic surface wave filters," *IEEE Trans. Microwave Theory Tech.* (*Special Issue on Microwave Acoustic Signal Processing*), vol. MTT-21, pp. 162–175, Apr. 1973.

Folded-Line and Hybrid Folded-Line Bandstop Filters

PAUL A. DUPUIS, MEMBER, IEEE, AND EDWARD G. CRISTAL, SENIOR MEMBER, IEEE

***Abstract*—The feasibility of a compact bandstop-filter geometry is demonstrated. The filter geometry is particularly suited for stripline and microwave-integrated-circuit (MIC) fabrications in that grounding is not required for any part of the filter. Hybrid geometries allow the filter designer increased flexibility in choosing a suitable shape factor without significantly affecting the filter's electrical characteristics. Folded-line filter geometries are suitable for narrow bandwidths (provided capacitive-coupled stubs are used) to wide-bandwidth applications. Experimental confirmation is presented.**

INTRODUCTION

The fabrication of microwave components and systems in stripline and microwave integrated circuits (MIC) has advantages to alternative realizations with respect to size, weight, and (usually) reproducibility. To minimize the number of substrates in a given system or subsystem, the number of components per substrate should be maximized consistent with isolation requirements between components. The compact bandstop-filter geometry described in this

Manuscript received April 12, 1974; revised June 26, 1974. The work reported in this short paper was conducted at McMaster University, Hamilton, Ont., Canada, and was supported by the National Research Council of Canada under Grant A8242.
P. A. Dupuis is with the Canadian Center for Inland Waters, Burlington, Ont., Canada.
E. G. Cristal is with the Hewlett-Packard Laboratories, Palo Alto, Calif. 94304.