module. It then summarizes each element of data in a chart that states if the acceptance specification has been met. The complete test operation for a single filter module requires 15 min.

V. Program Status and Conclusion

The primary application for the SENU for the DSCS calls for operational utilization of approximately 2000 of the filters in approximately 100 worldwide DSCS terminals. The efficiency of utilization of the available DSCS bandwidth is approximately 30 percent greater with the SENU filters than with alternative conventional group delayed equalized filters. A tactical variation of the SENU is also in the process of being produced for the U.S. Army Satellite Communication Agency.

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Programmable Frequency-Hop Synthesizers Based on Chirp Mixing

BARRY J. DARBY AND JOHN M. HANNAH

Abstract—**Frequency-hopped communication equipment require synthesizers capable of providing a large number (N) of discrete frequencies over a wide band. In typical systems N lies in the range of 50 to 10 000 and the required bandwidth ranges from 10 to 500 MHz.**

One technique for implementing a synthesizer is based on mixing chirp signals generated by impulsing SAW filters. Potentially, this method allows fast frequency hop generation over wide bandwidth (<500 MHz) with large numbers of selectable hop frequencies (N<4000). Furthermore, the hardware can occupy a small volume and dissipate low power compared with conventional synthesizers.

This paper examines the techniques and establishes likely parameter and performance bounds. Deleterious mechanisms are identified and their effects on CW spectral purity and fast frequency-hopped link error rate performance is discussed. Experimental results are presented for both a high performance modem, with N equal to 480 across a 96-MHz band and a recent development comprising the basic chirp synthesizer plus phased locked loop (PLL) to provide enhanced slow frequency hop and continuous-wave (CW) spectral purity.

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I. Introduction

SPREAD SPECTRUM modulation techniques provide the capability of low error rate communications in the presence of high cochannel interference levels [1]. This is achieved by modulating the transmitted low rate data with a characteristic signal possessing a bandwidth typically in the range 10 to 500 MHz. At the receiver, data are recovered at low input signal-to-"noise" ratios (SNR's) through cross correlation with a replica of the wide-band signal employed at the transmitter. During normal link operation, the predetection correlation process results in an enhancement of SNR (processing gain) determined to first order by the ratio transmitted bandwidth-to-data rate.

In frequency-hopping (FH) systems the available bandwidth is divided into a large number of contiguous subchannels. Bandspreading is achieved by transmitting successive constant duration pulses on pseudo-randomly selected subchannel carrier frequencies. The resulting wideband signal thus consists of a sequence of mutually orthogonal frequency-shift keyed pulses pseudo-randomly (PN) hopped over the total bandwidth. The many-to-one mapping of PN code states onto carrier frequencies is determin-

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istic and thus allows the necessary duplication of the FH sequence at a remote, synchronized receiver. Further, the achievable uniform occupancy of PN logic states results in the desired rectangular spectrum. In addition, an important feature of FH is the ability to incorporate, in the control logic, both programmable bandwidth (in discrete steps from a single carrier through to wide-band coverage) and programmable hop rate.

Important parameters for FH are the number of selectable frequencies (N), the total spread spectrum bandwidth (B), and the frequency hop rate. These three parameters effectively determine the ECCM capability of the system [1]. In general, N may range from ~ 50 to $>10\,000$, B may range from 10 to 500 MHz, and the hop rate vary from seconds to microseconds depending on the unique system requirements. Fast hopping is of interest to counter repeat jamming.

This paper is concerned with FH generation based on mixing two time delayed but partially overlapping linear frequency-modulated (FM) (chirp) waveforms. Implementation of this technique using SAW chirp filters was first pointed out by Atzeni *et al.* [2] and since then several previous papers [3]–[5] have explored the possibilities. The use of SAW here is particularly attractive since the range of T, B available [5] encompass the FFH requirement. Here, the important features of the technique, including the many mechanisms which have been identified as degrading the performance of an FH link, are discussed. Experimental results are presented for a phase locked loop addition to the basic synthesizer giving improved spectral purity for slow FH and CW modes of operation. Further, a complete fast FH modem covering a 96-MHz band in 480 subchannels is described in detail.

II. FH Generation Based on Chirp Mixing

The particular implementation discussed here is based on mixing two time-delayed overlapping chirp waveforms of the same numerical slope (μ) which have been produced by impulsing SAW filters. As shown previously [5], for perfect chirps, a constant frequency output can be obtained at either the sum or difference component of the mixer output. Selecting the sum frequency term, the instantaneous angular frequency is

$$\omega_s = (\omega_1 + \omega_2) - \mu_2 \tau + (\mu_1 + \mu_2)\tau. \qquad (1)$$

If $\mu_1 = -\mu_2 = \mu$ then the residual chirp term vanishes and the output frequency is linearly dependent on τ. Similarly, the instantaneous difference angular frequency is given by

$$\omega_d = |\omega_1 - \omega_2| + \mu_2 \tau + (\mu_1 - \mu_2)\tau. \qquad (2)$$

Choosing $\mu_1 = \mu_2 = \mu$ in this case will remove the chirp term and so give a similar result to the sum carrier frequency example.

In order to produce a constant hop dwell time (T_h), a time gate is applied to the mixer output and the range of τ is constrained to $\pm T_h = \pm T_0/2$, where T_0 is the chirp duration. Under these conditions [5] the number of orthogonal hop symbols (N) is maximized

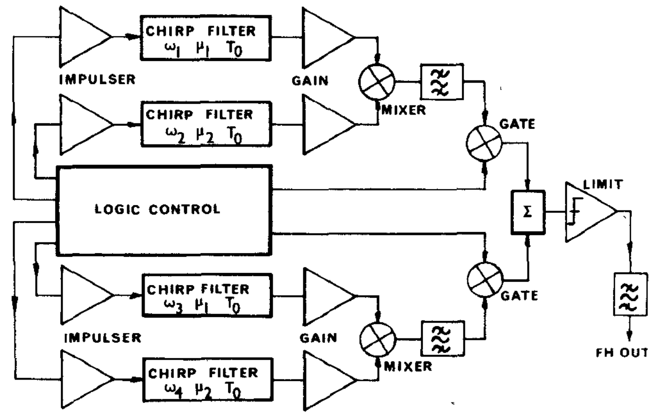


Fig. 1. Basic functional block diagram of a PN–FH chirp mixing synthesizer.

$$N = \tfrac{1}{2} B T_0 \qquad (3)$$

where B is the chirp bandwidth. The FH process can cover bandwidth B, and a 50-percent duty cycle output is obtained. Time interlacing of similarly produced waveforms will then give a continuous output.

Flexible operation is possible when the impulse timing to the SAW filters is under digital control with timing increments derived from a stable master clock. A minimum clock frequency of $B/2$ results conveniently in orthogonal FH symbols being obtained at delays (τ) equal to integral numbers of clock periods. Fig. 1 shows a generalized block diagram of the basic synthesizer. The digital control circuitry performs the function of providing the appropriate impulse timing signals required to generate a given frequency. All circuitry is driven synchronously from a stable master clock and consists basically of programmable counters which are preset by the input PN code. The output pulses, which are synchronous with the clock waveform, are fed to the appropriate impulse generator which produces a $\sim$40-V baseband impulse to excite the chirp filter. The digital control circuitry also produces the gate waveforms timed to allow interlacing of the two channels to obtain a continuous output signal.

Although not yet implemented, the following modifications to the basic configuration can be made.

1) FSK data modulation with deviation $\pm m$ channels for n hops per data bit ($n = 1, 2 \cdots$) can be included with a simple change to the digital control logic.

2) Instead of the rectangular waveform used to gate the signals in each channel a weighted, e.g., raised cosine, Hamming, etc., gate signal can be stored in ROM in order to reduce the spectrum splatter, caused by the sinc function spectrum envelope of each hop.

If it is desired to use interlaced pulses to obtain a continuous-wave (CW) output the hop length T_h must be an integral number of cycles of the synthesized frequency. This condition will be satisfied if each synthesized frequency is a harmonic of the frequency spacing, $1/T_h$. For the reasons given below, high purity i.e., <-60-dBc spurs are unlikely to be obtained through chirp mixing alone. (A following section describes a simple addition to the basic synthesizer which greatly improves signal purity in slow hop and CW modes.)

III. Factors Affecting FH Modem Performance

A number of factors influence the performance of either a fixed carrier or FH carrier communications link. These include the "quality" of the synthesized carrier both in transmitter and receiver, local oscillator instability, channel group delay variation, and inaccuracy of receiver tracking loops. Any of these factors may first predicate the choice between coherent and noncoherent signaling techniques, determine the minimum data rate, and ultimately limit the performance of the link.

The aim here is confined to identifying degradations in the quality of the synthesized carrier arising from inaccuracies in SAW device parameters and in peripheral circuitry. Characterization of the synthesizer performance involves investigation of many potential sources of error which, in general, affect: frequency accuracy and stability, carrier phase coherence, and spurious signal levels. These potential sources of error include: 1) impulse timing errors arising through fixed time offsets, clock instability, and impulse jitter; 2) chirp slope mismatch which results in a residual linear FM term, (1), (2); 3) chirp phase ripple which gives rise to a variable FM term depending on τ; 4) chirp center frequency errors which directly degrade frequency accuracy and invalidate the phase coherence condition; 5) gate edge timing errors which directly influence spurious levels; 6) relative amplitude, frequency, and phase imbalance between time interleaved sources which result in spurious modulation of the carrier; and 7) mixer spurii which arise through harmonic generation and result in increased "background" spurious level.

Items 2), 3), 4), 5), and 6) arise through the specific technique described here and apply to both FH and CW modes of operation. Aspects of item 1) arise in any direct synthesis approach. In addition, fixed time offsets result here in frequency offsets. These fixed offsets may arise in delay offsets in logic components, impulser circuits, group-delay differences between SAW devices, and group-delay differences in amplifier chains. In other direct synthesizers gate timing errors [5] only affect FH performance, here accurate gate timing is crucial for low spurious CW operation. Mixer spurii [7] arise with direct synthesizers and effectively determine the basic frequency plan of the filter banks.

The effect of temperature change merits especial mention. It is important to realize that positions of the spectral lines in the output are determined by the period of the gate waveform. Practically, this waveform is derived by synchronous division of the master clock. Thus although the amplitudes of the output spectral lines are a function of the chirp generation and mixing process their absolute position is not. In particular, the amplitudes of the spectral lines change because the instantaneous frequency generated by each of the interlaced synthesizer channels varies with temperature.

A number of factors could contribute to changes in the generated frequency as the temperature varies. In a well-designed synthesizer employing an ovened master clock, the limiting factor is likely to be the temperature characteristics of the SAW chirp filters. These devices are normally fabricated on ST-cut quartz to obtain the best temperature stability. Adopting the temperature data for ST quartz of Schulz and Holland it can be shown from (1) that for a sum-frequency synthesizer

$$\omega_{s\theta}=\frac{1}{\beta}\left(\omega_{1_{\theta_0}}+\omega_{2_{\theta_0}}\right)-\frac{1}{\beta^2}\left(\mu_{\theta_0}\tau\right) \tag{4}$$

where the subscript θ refers to the value of the parameter at temperature θ, θ_0 is the turnover temperature ($\sim 21°C$), and $\beta=(1+\alpha(\theta-\theta_0)^2)$, ($\alpha \sim 32\times 10^{-9}\,°C^{-2}$). A consideration of (4) shows that for practical synthesizers, the first term will have the dominant effect (since normally $\omega_1+\omega_2 \gg \mu\tau$). This equation reflects the familiar parabolic shape expected for ST-cut quartz and results in, for example, a $\sim$80-ppm change in frequency for a 50°C variation from θ_0.

In CW mode, this shift in frequency will initially produce increased spurious signal levels due to the shift in the sinc function nulls and phase reinforcement of the spurious outputs from each channel.

For fast frequency hopping (FFH) operation, the effect on hop frequency due to temperature change of the SAW devices can be made small compared with the hop bandwidth and results in a small implementation loss on demodulation.

Computer simulations were performed for CW operation of the synthesizer. The maximum spurious level obtained was computed for a range of error values with each of the mechanisms: 2), 3), 4), 5), and 6) incorporated in turn. This is a very stringent test, degraded CW performance due to any of these mechanisms does not necessarily imply inadequate performance in noncoherent FH links. Spurious levels were found by computing the spectrum of the periodic time domain signal using the fast Fourier transform.

Specifically, Fig. 2 gives the maximum spurious (spur) level as a function of: 1) slope mismatch between up and down chirps, $\Delta\mu/2\pi$ (MHz/μs), giving rise to a periodic chirp signal; 2) frequency error, Δf(MHz) on each synthesizer channel, which can arise from mechanisms 1), 4), and temperature change and violates the phase coherence condition; 3) phase imbalance, $\Delta\phi$(degrees), between synthesizer channels; 4) gate pulse-length variation, ΔT_g (nanoseconds); and 5) chirp phase ripple, $\Delta\theta$ peak (degrees), assumed to be a one cycle sinusoidal departure from the expected time-domain quadratic-phase response.

In addition, Fig. 2 shows the effect of amplitude imbalance between synthesizer channels.

It is evident from Fig. 2 that the synthesized signal is

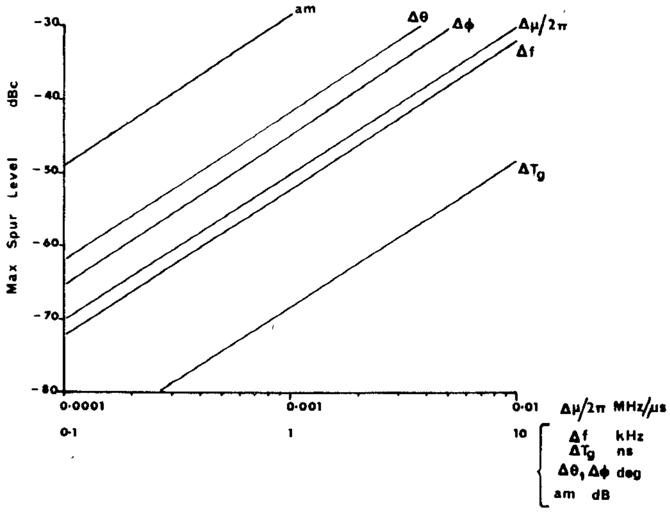


Fig. 2. Computed maximum spurious levels (2.5-μs hops).

least sensitive to errors in the gate duration (ΔT_g). Spur levels of -60 dBc can be achieved with attention to the design of the RF switch and its driving circuit.

Clearly, achievement of this level of CW performance in respect of matching the chirp slope and the device center frequency accuracy is beyond the current state of the art. (Minimum slope mismatch is predicted to be obtained when the upchirp is obtained by spectral inversion of the downchirp.) Likely performance bounds for FH and CW synthesis are discussed below.

The final source of spurious to be considered here is due to the generation of harmonics and consequent intermodulation products in double balanced mixers. One particularly serious component is obtained from the intermodulation of the third harmonic of the "local oscillator" (LO) with the input "RF." The level of spurious obtained is only ~ 15 dB down from the desired signal for CW inputs. In many other situations, this spurious signal can be removed by appropriate frequency filtering. However, as shown below, the existence of this signal imposes a further restriction on the design of synthesizers for sum frequency outputs. In this application, one chirp signal is amplified to a high level ($+7$ dBm) and applied to the "LO" port on the mixer. The other chirp signal is maintained at a relatively low level (~ 0 dBm), to avoid enhancement of other harmonics, intermodulation ($3\times$LO$-$RF) spurious is at: $3\omega_1-\omega_2+4\mu t$ and, the desired frequency component is a carrier at: $\omega_1+\omega_2-\mu\tau$. Both signals exist for time $(T_0-\tau)$. In order to avoid the spurious appearing in the output bandwidth B, a simple analysis shows that

$$|\omega_1-\omega_2|>1.25B.$$

This restriction points to the use of sideband inversion to produce the upchirp signal spaced $\pm 1.25B$ from the center frequency of the downchirp.

The error rate performance as a function of input SNR for an FH digital communications link depends on a number of factors. Chief among them are the coherence and the accuracy of match between transmitted FH signal and receiver FH reference waveforms. Coherence allows predetection integration over more than one hop period thus giving a processing gain greater than the 10 $\log_{10} N$ (dB) improvement offered by noncoherent links. However, if the elementary hop signals produced at the transmitter and receiver are mismatched then optimum output SNR will not be obtained.

Here, two important sources of error are a constant frequency offset and differences in the residual linear FM component on both signal and reference. Since receiver tracking loops serve to minimize frequency offsets, the prime problem is the difference in residual FM on the cross-correlated waveforms. Two cases of interest arise: 1) all FH links in the system use the same equipments; 2) interoperability is required with equipment employing other FH synthesizer techniques.

In the first case, matched filter theory allows an arbitrary residual FM component provided that it is identical on both signal and reference waveforms. The degree of match obtained here is, therefore, restricted by the replication accuracy of SAW devices. In the second case, the residual FM component must be minimized. Here the degree of match is determined both by design accuracy and by replication accuracy.

It has been shown [5] that for less than 1-dB mismatch loss the maximum chirp BT product is of the order of 3000 when replication and design accuracies of 2 parts in 10^4 can be achieved.

IV. Fast Frequency Hopping Modem

In common with previously reported designs [5], the synthesizer described here makes use of sum frequency operation and gives a continuous time domain output instantaneously producing a fast FH (FFH) carrier under control of a PN code.

The current experimental hardware is configured to generate the FFH carrier in two separate bands each 48 MHz wide. Frequencies between 306 and 354 MHz are covered by the lower band while the upper band occupies 356 to 404 MHz. Two pairs of SAW chirp filters, one pair associated with each band, are employed; details of parameters are given in Table I. Thus the output frequency hops alternately between upper and lower bands with PN selection from a set of 240 subchannels in each band. It would be difficult even for sophisticated jammers to exploit this systematic feature. To be effective, a repeater jammer would have to be located well within an ellipsoid of revolution whose boundary is determined by the condition that the direct path delay between receiver and transmitter and the indirect path delay, via the jammer, differ by the hop duration. For example, with transmitter and receiver separated by 50 km, the ellipse semi-minor axis is only ~ 6 km for 5-μs hops. In practice, this small spatial region can be reduced by using antennas with narrow radiation patterns. For other types of jammer, the exact location of the victim receiver and the transmitter must be known; a task made more difficult with multiuser nets.

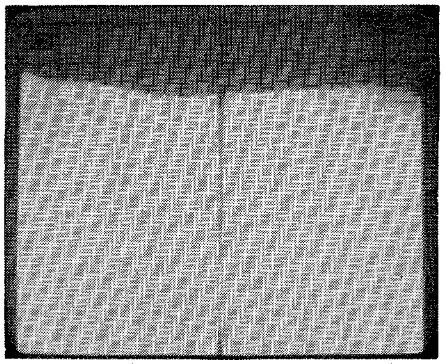


Fig. 3. Output FFH spectrum (scales: 1 dB/div vertical and 10 MHz/div horizontal).

TABLE I
CHIRP FILTER PARAMETERS

Centre frequency (MHz)		
Lower band:	165.0	Up and down chirps at same f_0 for each band
Upper band:	190.0	
Chirp duration: -20 dB points (μsec).	10.0	
Chirp slope: (MHz/μsec)	$\pm$ 4.8	(design)
Measured slopes:		

f_0 (MHz)	$+\mu/2\pi$	$-\mu/2\pi$
165	4.787, 4.786	4.792, 4.789
190	4.781, 4.783	4.788, 4.790

Departure from quadratic phase:	<2.0° rms down chirps; <4.0° rms up chirps
Geometry:	double dispersive, in-line with Fresnel ripple compensation on ST, X quartz

For this modem, frequency selection is done using an ROM to perform the many-to-one mapping of 10-bit states of a 15-stage PN code generator onto counter states. Thus the delays are achieved by continuously decrementing these synchronous counters with a 24-MHz master clock signal. The resulting mutually delayed "underflow" pulses are used to trigger avalanche impulse generators. The resulting packaged hardware is compact, measuring only 300$\times$140 $\times$70 mm and has low power consumption— 13 W. It is predicted that at least a factor of two reduction in size and power consumption can be made.

Fig. 3 shows the spectrum of the FH carrier where the 1-dB/vertical division scale highlights the flatness of the spectrum produced by this technique. Further, back to back coherent demodulation tests indicate the accurate frequency duplication obtained across the whole bandwidth by two synthesizers. This has been confirmed by measuring spot frequencies; less than 20-kHz error between synthesizer outputs was found. In the FFH modem, the data rate and hop rate are equal. A 20-kHz offset results in only $\sim$0.15-dB loss in SNR for the 5-μs hop duration.

An experimental link is currently being tested. At the transmitter, binary FSK data modulation is accomplished by the "phasing" single-sideband circuit technique. This approach was favored so that experiments concerned with varying the ratio of 200-khop/s rate to data rate could be performed. At the receiver, synchronization acquisition is obtained by one-way serial search through the receiver's region of time uncertainty. This is simply carried out by deleting master clock pulses. After obtaining coarse synchronization, a delay-lock tracking loop maintains signal and receiver FH reference in alignment.

In this method, two separate correlators are supplied with reference signals mutually delayed by one hop duration T_h. The outputs of the correlators are combined to produce a discriminator response. Tracking is achieved at the signal-reference delay corresponding to $\frac{1}{2}T_h$.

For the current modem, the offset delays of 2.5 and 5.0 μs were conveniently obtained using a wide-band SAW tapped delay. The 100-MHz bandwidth was obtained with identical dispersive input and output transducers, TB product 50, arranged to give nondispersive delay on a lithium niobate substrate. The insertion loss to each tap was approximately 30 dB, small variations ($<$2 dB) in the output levels were reduced by limiting amplifiers. Spurious time

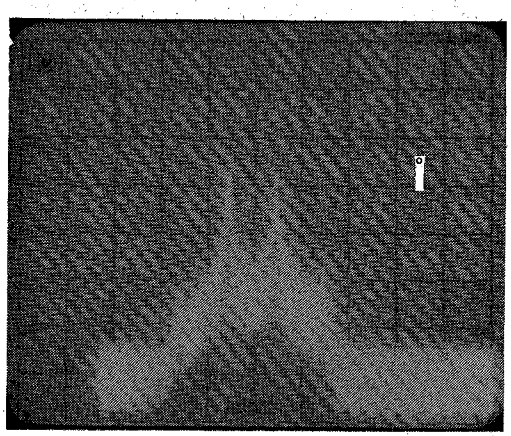


Fig. 4. Spectrum of the dehopped FSK signal centered on 60 MHz (scales: 10 dB/div vertical and MHz/div horizontal).

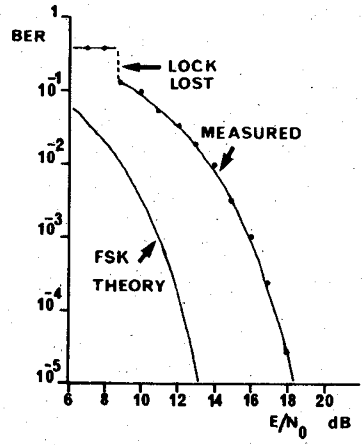


Fig. 5. FFH modem performance bit error rate versus E/N_0.

domain echoes were 30 dB down on the delayed outputs and gave negligible perturbations to the discriminator characteristic.

Fig. 4 shows the spectrum of the dehopped binary FSK signal centered on 60 MHz, where 1 bit is transmitted per frequency hop interval. This signal was taken at the output of the cross correlator, which has a 4-MHz bandwidth. Subsequent signal processing comprises narrow-band IF filters centered at the mark and space frequencies in the first stage of the conventional FSK receiver.

Data demodulation is accomplished by the standard method of comparing mark and space integrate and dump filter outputs at the end of the data bit intervals. Fig. 5 shows a preliminary bit error rate measurement made for band-limited white Gaussian noise interference for the case where 1 bit is transmitted per frequency hop interval with binary FSK signaling. The performance depicted is obtained after upconversion to a microwave carrier and subsequent downconversion with spread spectrum and data demodulation and includes the effects of automatic gain control (AGC) and tracking loop performance. The deviation in measured performance from theory is attributed mainly to correctable deficiencies in implementing the conventional FSK receiver and in the AGC loop. It is estimated through separate measurements on the FSK demodulator that the implementation loss due to the SAW synthesizer and correlator amount to <2 dB.

V. Reduction of Spurious Levels in CW Mode

The level of spurious signals, when generating CW waveforms, has been shown to be very dependent on the accuracy of design and fabrication of the SAW filters used. The best achieved performance, to date, was a spurious signal level of -32 dBc [5] which is considerably above the <-60 dBc desired in some applications.

Fig. 6 shows the block diagram of a tracking filter module which has been designed to achieve a low level of spurious signals for CW or slow hop modes, while retaining the capability of switching the output to the chirp mixing synthesizer for FFH. This consists of a phase locked loop (PLL) which tracks the FH synthesizer output. The loop filter rejects the unwanted spurious signals from the hopper, given an adequate carrier to spurious ratio, producing a "cleaned-up" output waveform from the voltage-controlled oscillator (VCO). The input mixer and the divider in the feedback loop were required because the

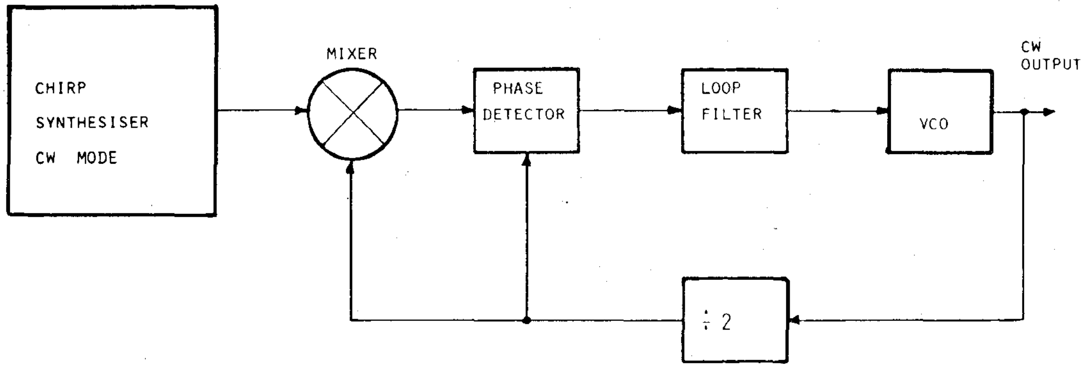
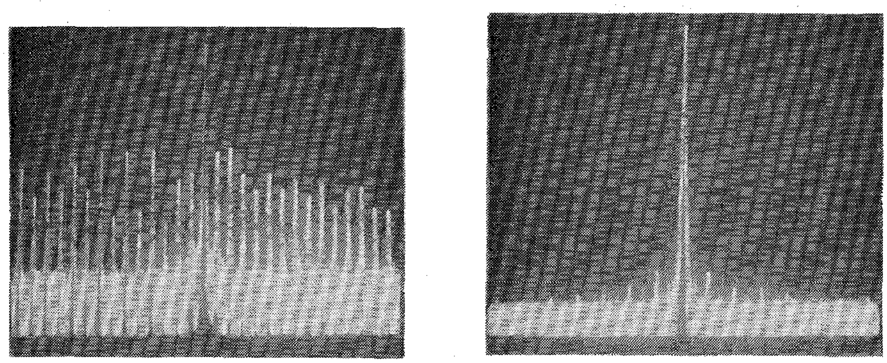
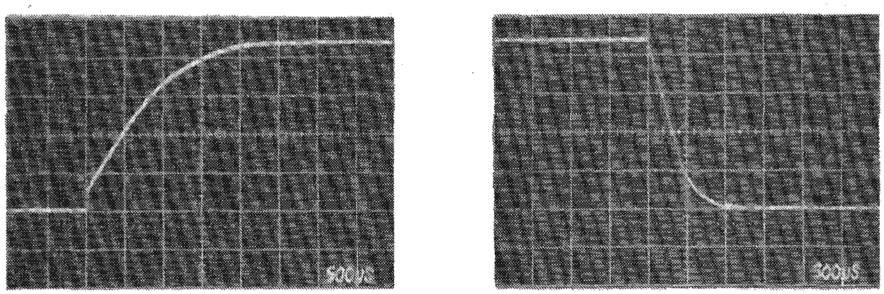


Fig. 6. Block diagram of tracking filter.



(a) (b)

Fig. 7. (a) Output spectrum in CW mode (scales: 10 dB/div vertical and 500 kHz/div horizontal). (b) Output spectrum after passing through tracking filter (scales: 10 dB/div vertical and 500 kHz/div horizontal).



(a) (b)

Fig. 8. VCO control line voltage (scales: 2 V/div vertical and 500 μs/div horizontal). (a) 107.5- to 132.5-MHz jump. (b) 132.5- to 107.5-MHz jump.

prototype module was constructed using standard emitter coupled logic (ECL) components to cover the frequency range from 107.5 to 132.5 MHz. This tracking filter circuit occupies a small area ($\sim 125 \times 100$ mm) of printed circuit board.

Fig. 7(a) shows the output spectrum from a prototype FH synthesizer [5] operating in CW mode at 120 MHz. The spectrum in Fig. 7(b) is that observed at the output of the tracking filter module under the same conditions. This shows clearly the excellent rejection of spurious signals achieved by this unit which consistently gives better than -60-dBc spur level over the whole hopper bandwidth. Slow FH operation with a clean output signal is possible using this circuit provided the hop duration is much longer than the switching speed of the loop. Fig. 8(a) and (b) shows that for a 25-MHz frequency jump the average VCO switching speed is 2 ms. A detailed examination of the phase of the output carrier would determine a lower bound on data rate in this mode.

VI. Conclusions

It has been shown that the technique of mixing chirp signals produced by impulsing SAW filters enables FFH generation over a wide bandwidth with large numbers of selectable hop frequencies in simple, compact, low power hardware and provides a new capability in this area. The effects of imperfections in the SAW chirp filters on FH modem performance and CW generation have been discussed and it has been shown that although low spurious CW performance is beyond the current state of the art, low error-rate FFH modem operation is possible with available device designs.

In addition, a simple tracking filter module has been described which is capable of giving low spurious (-60 dBc) CW signals with acceptable switching speed. This module allows operation on fixed or slow hopping carriers in addition to the FFH capability. The combination of the two modules is not intended to compete with established synthesizer techniques for applications involving only slow FH.

Acknowledgment

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Acoustoelectric Convolver Technology for Spread-Spectrum Communications

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***Abstract*— Acoustoelectric (AE) convolvers for spread-spectrum communication applications are described with input bandwidth capacities to 200 MHz. These devices offer an unique combination of large processing gain, high dynamic range, small size and weight, and low drive power requirements. The programmable feature of convolvers allows the encoding waveform to be changed from bit-to-bit, thereby providing resistance to repeat jamming and enabling secure communications.**

The basic concepts of a convolver-based spread-spectrum communications system are reviewed, current convolver capabilities are discussed, and projections are made for future device performance. Deviations from nonideal convolver performance are considered. Special techniques which must be used in the system implementation and evaluation of convolvers are described, and the performance level achieved in a state-of-art convolver subsystem is given.

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I. Introduction

The convolver [1] can function as a programmable matched filter for wide-band spread-spectrum waveforms. The programmable feature of convolvers allows the encoding waveform to be changed from bit-to-bit, thereby providing resistance to repeat jamming and enabling secure communications. Since the matched filter function is determined electronically in a convolver, this device does not suffer from the temperature related phase problem of a conventional surface-acoustic-wave (SAW) matched filter with a comparable time-bandwidth product. A dual convolver [2] provides a convenient method of demodulating

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