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A Flexible, HF Sweep-Frequency Sounder With Submicrosecond Resolution

by

L. L. Peden
J. M. Lomasney
R. B. Fenwick

November 1968

Technical Report No. 140

Prepared under
Office of Naval Research Contract
Nonr-225(64), NR 088 019, and
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RADIOSCIENCE LABORATORY

STANFORD ELECTRONICS LABORATORIES

STANFORD UNIVERSITY · STANFORD, CALIFORNIA

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ABSTRACT

A frequency-modulation, continuous-wave HF ionospheric sounder ("Chirp I") was introduced in 1964. This equipment provided a considerable increase in system sensitivity over commonly used pulse sounders at a time-delay resolution of 10 μsec. Experience with this equipment revealed a need for still greater time-delay resolution capability, plus easily selectable sweep-frequency limits and rates, and an ability to sweep with either positive or negative slopes.

In 1966, a second generation equipment, "Chirp II," was developed. This report describes the differences between it and the "Chirp I" equipment. The salient characteristics of "Chirp II" are time-delay resolution of 1 μsec or less, great flexibility in selection of sweep limits and sweep rates, and a choice of several sweep formats. The sweep phase discontinuities have been reduced to a level more than adequate for most ionospheric sounding work, but further reduction is needed in some applications.

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I INTRODUCTION

A. History

A frequency-modulation, continuous-wave HF ionospheric sounder ("Chirp I") employing a precise, direct-synthesized waveform was introduced in 1964 by Fenwick and Barry [Ref. 1]. Advantages over normally used pulse techniques included high time-delay resolution, increased immunity to narrowband interference, and unity peak to average power ratio.

The Chirp I sweep generating equipments have been used for a variety of experimental tasks, including vertical-incidence, oblique-incidence, and backscatter sounding [Ref. 2], and extraterrestrial radar measurements. A survey article by Barry and Fenwick describing sounding with synthesized frequency sweeps was published in the July 1965 issue of Hewlett-Packard Journal [Ref. 3].

Oblique soundings have been made over a number of paths of widely different lengths. Soundings were made over a 2000-km path, with transmissions originating at Lubbock, Texas, and received at Stanford. Most soundings were made in the frequency range 4 to 32 MHz, using power levels between 20 mW and 100 W.

The first high-resolution oblique soundings in the MF broadcast band were also accomplished over the Stanford-Lubbock path [Ref. 4]. These soundings were made at night in the 0.6- to 2.0-MHz frequency range using radiated power levels varying from 25 to 250 W. A notable feature of these experiments was the fact that interference from fixed broadcast stations was not troublesome, while a time-delay resolution of 10 μsec was achieved.

Oblique soundings over an 8000-km path were conducted by Murphy and Fenwick in 1965 [Ref. 5]. Sounding transmissions originating at Pahoa, Hawaii, were received at South Dartmouth, Massachusetts. Sweep rates of 25, 100, and 1000 kHz/sec were used with power levels from 1 W to 10 kW. Sounding records were compared with those made by an oblique pulse sounder operating alternately with the Chirp equipment over the same path.

Chirp I equipments were operated as an oblique sounder over a 9500-km path from Okinawa to Italy from February to August 1966. Ionograms made during July 1966 were analyzed to determine the extent to which oblique ionograms can reveal ionospheric structure on such long paths [Ref. 6].

"Around-the-world" (40,000-km path) soundings were made using Chirp I equipments with 100-W signals (6-32 MHz) transmitted eastward from Stanford and received from the west at Pescadero, California, at a sweep rate of 25 kHz/sec.

Vertical-incidence soundings were made over the 2- to 10-MHz frequency range, and at a frequency-sweep rate of 25 kHz/sec. The receiver was located at Pescadero, California, and the transmitter at Stanford--a site separation of 20 miles.

Backscatter soundings were made in the 6- to 32-MHz frequency range, with transmission originating at Stanford and received at Pescadero, California. Average transmitted power was 100 W.

Radar measurements of the total cislunar electron content were made in 1964 by Howard and Eshleman [Ref. 7]. The measurements were accomplished by determining the difference in time delay between 25 and 50 MHz moon-reflected signals. The Chirp I equipment was used to generate the transmitter and receiver frequency-sweep waveforms.

B. Need for Improved Sweep-Generating Equipment

Although considerable new experimental work was made possible by the Chirp I equipments, several equipment limitations were felt. The chief shortcoming, and the motivating reason for designing the Chirp II equipment, was the time-delay resolution limitation of 10 μsec (due to phase discontinuities occurring at 100-kHz intervals). Other disadvantages included the lack of easily selectable sweep limits and sweep rates, and the ability to sweep only in the direction of increasing frequency.

The equipment to be described in this report has improved resolution capabilities, and features to make operation more flexible and convenient.

II SYSTEM REQUIREMENTS

In early 1966, development commenced on "Chirp II." Characteristics of Chirp I, desired Chirp II characteristics, and achieved "end of development" results are listed in Table 1.

A. Resolution

The new sounder, of which the chirp generator was to be a fundamental component, was designed to measure the time-delay spread of ionospherically propagated HF radio signals. It was estimated that F-layer reflected signals would not exhibit a time-delay dispersion of less than 1 μsec; therefore, a 3-dB resolution of 1 μsec (or 1-MHz coherent bandwidth) was taken as a primary design criterion.

B. Synthesizer Description

In order to understand the discussion of problem areas in subsequent sections, it is necessary to be familiar with the digital synthesizer used, which generates the FM-CW sweep signal.

The synthesizer is a Hewlett-Packard Model 5100,* which is capable of producing an output frequency in response to decimal digital commands, set in either manually or by remote digital signals. The synthesizer is modified so that it can respond to digital commands from the Sweep Generator Control at the speed necessary to generate the sweep waveform.

The synthesizer is of the direct-synthesis type, operating on the NBS principle [Ref. 8] whereby frequency increments as small as desired are generated from a few base frequencies by repeated addition and division. Figure 1 shows a front view of the synthesizer, and Fig. 2 is a block diagram of a typical decade of the synthesizer, illustrating the decade division scheme used.

A signal from the previous decade, in the 3.0- to 3.1-MHz frequency range, goes to the first mixer at input "A," where it is mixed with the "B" signal, a fixed 24 MHz. The 27.0- to 27.1-MHz output is mixed again

---

*Manufactured by Hewlett-Packard, Palo Alto, Calif. 94304.

Table 1

SYSTEM CHARACTERISTICS COMPARISON: CHIRP I-CHIRP II

CHARACTERISTIC	CHIRP I	DESIRED: CHIRP II	ACHIEVED: CHIRP II
Resolution	10 μsec	1 μsec	~ 500 nsec
Coherent Bandwidth	100 kHz	1 MHz	~ 10 MHz
Sidelobes, ≤ 100 msec from main ramp (at 1 MHz/sec)	-20 dB	-20 to -35 dB	greater than -20 dB
Sidelobes, > 100 msec from main ramp (at 1 MHz/sec)	-15 dB	-30 dB or better	-30 dB
Phase accuracy of 3.0-3.9 MHz lines in HP 5100 at significant switching times	$\leq \pm 18°$	$\leq \pm 10°$	$\leq \pm 7°$
Phase accuracy of 30-39 MHz lines in HP 5100 at significant switching times	Random	As good as practical	$\leq \pm 15°$
Selectable sweep rates	25, 50 and 100 kHz/sec	25, 50, 100 and 1000 kHz/sec	
Sweep limit selection	Resolder limit lines to desired digit	Switch selectable to nearest 100 kHz throughout 0-50 MHz range	
Selectable sweep formats	Positive sweep slope only	Positive or negative sloped sweeps or combinations of both	
Transfer from sweep frequency to fixed frequency mode and vice versa	Change requires removal of two cables and actuation of 10 switches	Change by actuating one switch	

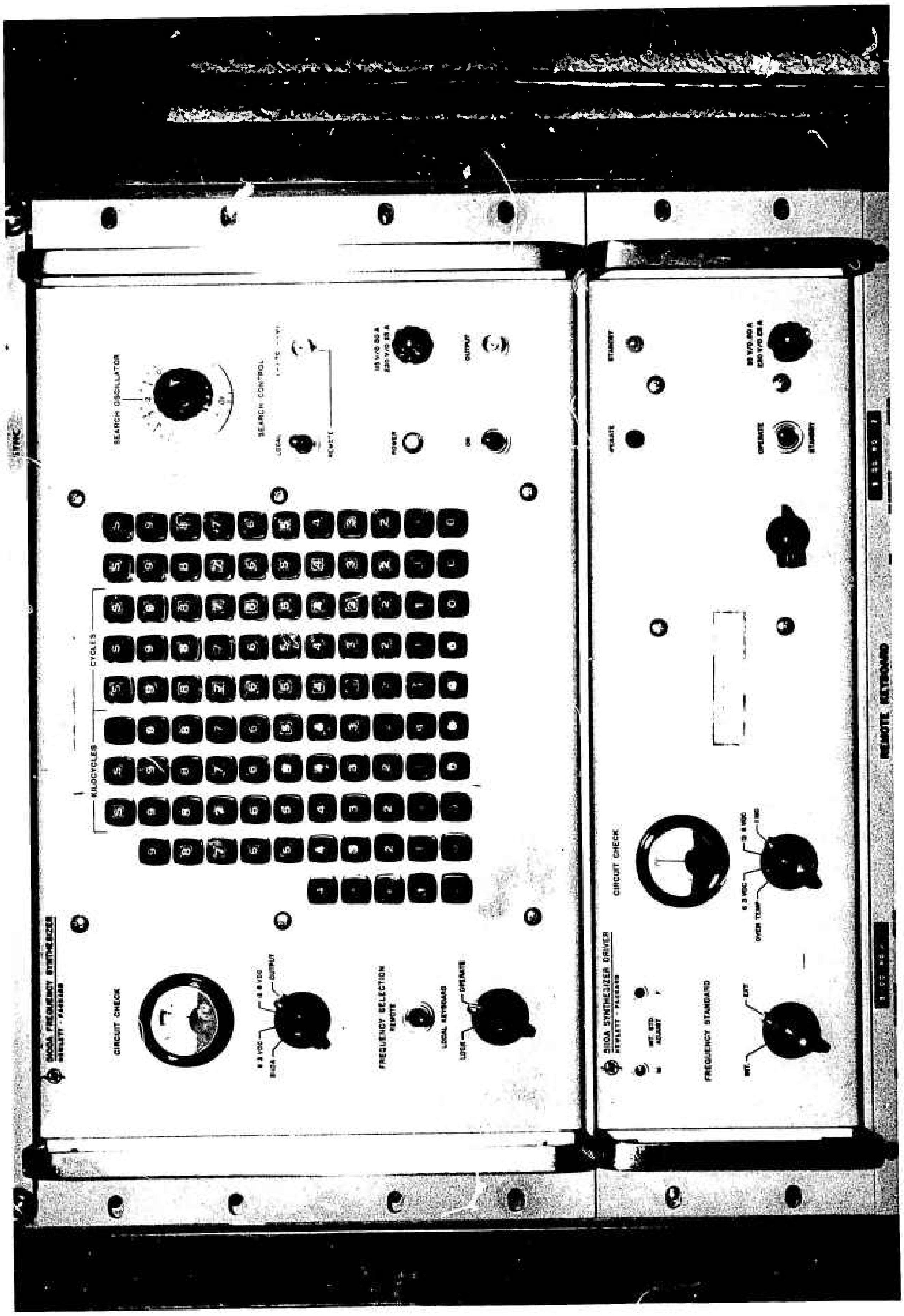


Fig. 1. THE HEWLETT-PACKARD 5100A FREQUENCY SYNTHESIZER.

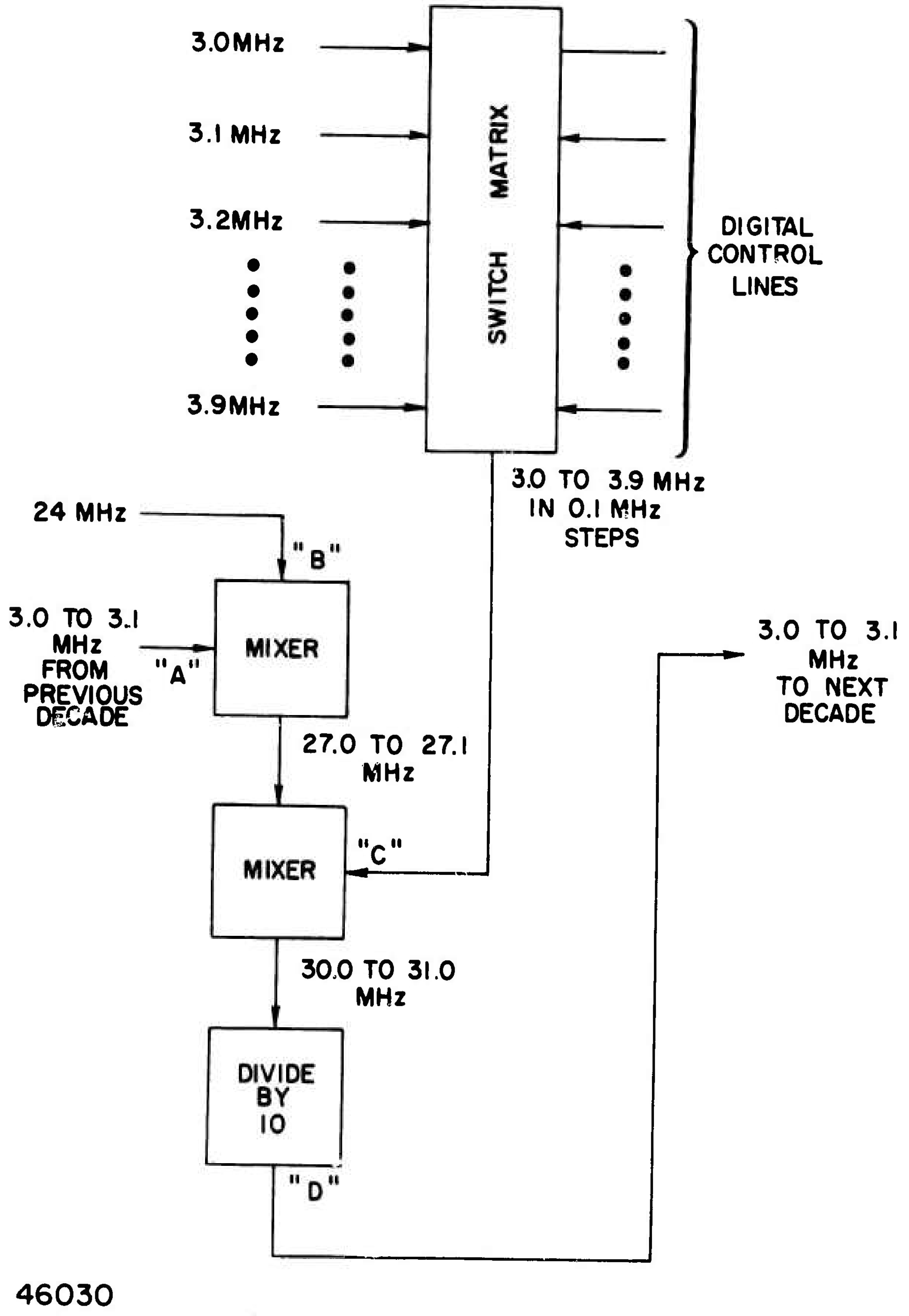


Fig. 2. BLOCK DIAGRAM, TYPICAL DECADE OF HEWLETT-PACKARD 5100A SYNTHESIZER.

with the "C" input, one of the ten signals 3.0, 3.1···· to 3.9 MHz selected by the switch matrix. The sum output, in the 30.0- to 31.0-MHz range, is then divided by ten to yield another 3.0- to 3.1-MHz signal to drive the next decade. Filters, not shown in Fig. 2, are provided following each mixer and divider, so that only the desired signal passes to the next stage.

Suppose the signal at "A" is adjustable in frequency with a smallest increment of 10,000 Hz (by selection of any one of the signals in its switch matrix); then the smallest increment in the output signal "D" is 1000 Hz, because of the division by ten. Repetition of this process can make the smallest frequency increment as small as desired. In the standard 5100A synthesizer, seven such decades are used to provide all frequency steps from 10 kHz down to a smallest increment of 0.01 Hz; an eighth decade, having no divider, provides 100-kHz increments. Thus any frequency within a 1-MHz range in steps of 0.01 Hz can be obtained by energizing the proper digital control lines to the seven (3.0- to 3.9-MHz) switch matrices. The 1-MHz and 10-MHz increments are obtained by choosing one frequency in the 30-, 31-····39-MHz range and one in the 350-, 360-····390-MHz range in later mixing processes. Again, these frequencies are chosen by control signals applied to switch matrices.

It is apparent, then, that a frequency ramp can, in principle, be generated by the synthesizer, if the proper sequence of digital control signals is applied to the switch matrices controlling the various frequency digits. It is necessary, however, to insure that the digital switching occurs at the proper times, so that the phase discontinuities between successive frequency signals are minimized, for reasons to be discussed.

Since frequency steps of 1 Hz were sufficiently small for the proposed use, and since the space was needed for other reasons, the two decades generating the least significant digits (0.1 Hz and 0.01 Hz) were not needed and were removed from the synthesizer used in Chirp II.

C. Sidelobe Levels and 3.0- to 3.9-MHz Synthesizing-Frequency Alignment

In the FM-CW sounding technique, sidelobes are defined as spurious-frequency signals which sweep simultaneously with the main frequency ramp but are offset by discrete frequency intervals on either side of the fundamental spectral component. Sidelobes which are close to the frequency of the desired ramp degrade the sounder's capability of resolving modes, and therefore must be suppressed to whatever extent is possible, consistent with reasonable cost. Sidelobes spaced at wider intervals, such as 100 kHz and 1 MHz from the frequency sweep fundamental component, should also be suppressed, to avoid unnecessary interference to other users of the radio spectrum.

Sidelobes arise in the sweep-frequency synthesizer primarily as a result of phase modulation (PM). This modulation is caused by phase differences at times of switching between the 10 base synthesizing frequencies 3.0, 3.1, 3.2····3.9 MHz and by time delays encountered in the mixer dividers. As the frequency ramp is produced, synthesizing frequencies 3.0, 3.1····3.9 MHz are successively selected at 25-, 50-, or 100-kHz/ sec rates. The phase discontinuities introduced at the switching times appear in the output signal (although divided in magnitude by consecutive factors of ten as they originate in decades less significant than the 10^5 decade). One therefore desires to minimize the phase error between the base synthesizing frequencies at the time of switching. A reasonable phase-alignment design goal for Chirp II was considered < 10 degrees. This, it was expected, would result in worst sidelobes 20 to 35 dB below the desired ramp (assuming decade time-delay problems were overcome, see below).

Time delays in the synthesizer mixer/dividers, and especially the time delay caused by filter circuits in the last three mixer/dividers prior to the UHF section of the synthesizer, give rise to phase errors in the output frequency sweep that are comparable in magnitude to those possible in switching unphased 3.0- to 3.9-MHz lines. These delays were not investigated in Chirp I development, but their effect on attainable time-delay resolution and sidelobe levels was the most serious problem encountered in Chirp II development. Thus means for reducing and/or compensating for the delays were sought.

D. 30- to 39-MHz Phase Continuity

By minimizing the phase discontinuities when switching between the 3.0- to 3.9-MHz base frequencies, as mentioned in the previous section, phase coherence over a 1-MHz increment of the sweep would be obtained. If the phases of the 30- to 39-MHz synthesizing frequencies (applied to the second mixer in the UHF section of the synthesizer) could also be made phase-identical at their prescribed switching times, coherence would be extended to a 10-MHz sweep increment.

In developing Chirp I, alignment of the ten higher-frequency lines was not considered, since the frequency synthesizer mixer/divider time-delays limited resolution to approximately 10 μsec. Because Chirp II resolution was not expected to be limited by phase errors at 100-kHz switching points, investigation of 30- to 39-MHz phase error and of means of reducing it was justified.

E. Sweep Limits and Rates

A major inconvenience encountered with Chirp I was the inability to change sweep frequency limits and sweep rates quickly. It was desired that Chirp II sweep limits should be controllable from decade thumb-wheel switches located on the front panel of the Sweep Generator Control. Limits controllable to the nearest 100 kHz were considered sufficient. Sweep rates of 25, 50, 100, and 1000 kHz/sec, selectable on the Sweep Generator Control panel, were also desired.

F. Selectable Sweep Formats

Versatility of sweep format was considered a necessary design criterion which implied use of decade counters (for control of synthesizing frequencies) that could operate in forward and reverse modes. Furthermore, the counters would be required to change direction in an amount of time that was small compared with the fastest incremental output frequency step. With quickly reversible counters, several sweep formats using combinations of positive and negative slope could be conveniently controlled by a few switches on the Sweep Generator Control front panel.

G. Fixed-Frequency Operation

In experience with Chirp I, it was found desirable to operate the exciter at a fixed frequency for transmitter tuning, receiver calibration, etc. With Chirp I, operation of the synthesizers in fixed-frequency mode required that two cables be disconnected and that several switches be actuated, including a frequency-selection button for each decade on the front panel of the synthesizer. In order to return to the sweep-frequency mode, all cables had to be replaced and frequency selection buttons had to be disengaged. In Chirp II, it was desired to change quickly and conveniently from sweeping to fixed frequency mode and vice versa.

III SYSTEM DESIGN AND DEVELOPMENT

A. General

Figure 3 shows an overall block diagram of the Chirp II linear sweep generator. It consists of three major parts:

(1) an HP 5110A Frequency-Synthesizer Driver,

(2) an HP 5100A Frequency Synthesizer,

(3) a Sweep Generator Control.

The Frequency Synthesizer Driver furnishes the various fixed frequencies required by the Frequency Synthesizer itself. The operation of the synthesizer was described briefly in Section II-B; it produces a directly synthesized output signal whose frequency is controlled by digital voltages in proper sequence. The frequency synthesizer was modified, as described in the following sections, to minimize noise and phase discontinuities in the output waveform and to obtain simplified switching between sweep-frequency and fixed-frequency modes of operation.

The Sweep Generator Control provides the sequentially-switched digital control voltage to the synthesizer. Figure 4 is a view of the front panel of the Sweep Generator Control unit.

B. Spurious Sweep-Frequency Components

1. Ideal Frequency Staircase

Consider an ideal, positive-slope, staircase frequency sweep. It is of interest to calculate the spurious signals (the "sidelobes" referred to above) that arise from the stepped-frequency characteristic of the synthesized signal. The frequency difference between staircase and ideal linear ramps is a negatively sloped sawtooth with peak-to-peak frequency error f_i and period t_i, the frequency-step period. Integration of the frequency error gives the phase error, a series of parabolic curves with period t_i. The peak-to-peak phase error, $\emptyset_e$, is equal to $\frac{\pi}{4} f_i t_i$.

The importance of this phase error is most easily seen by imagining that the ideal linear sweep is represented by a phasor. The synthesized phasor oscillates about the ideal linear phasor, and for

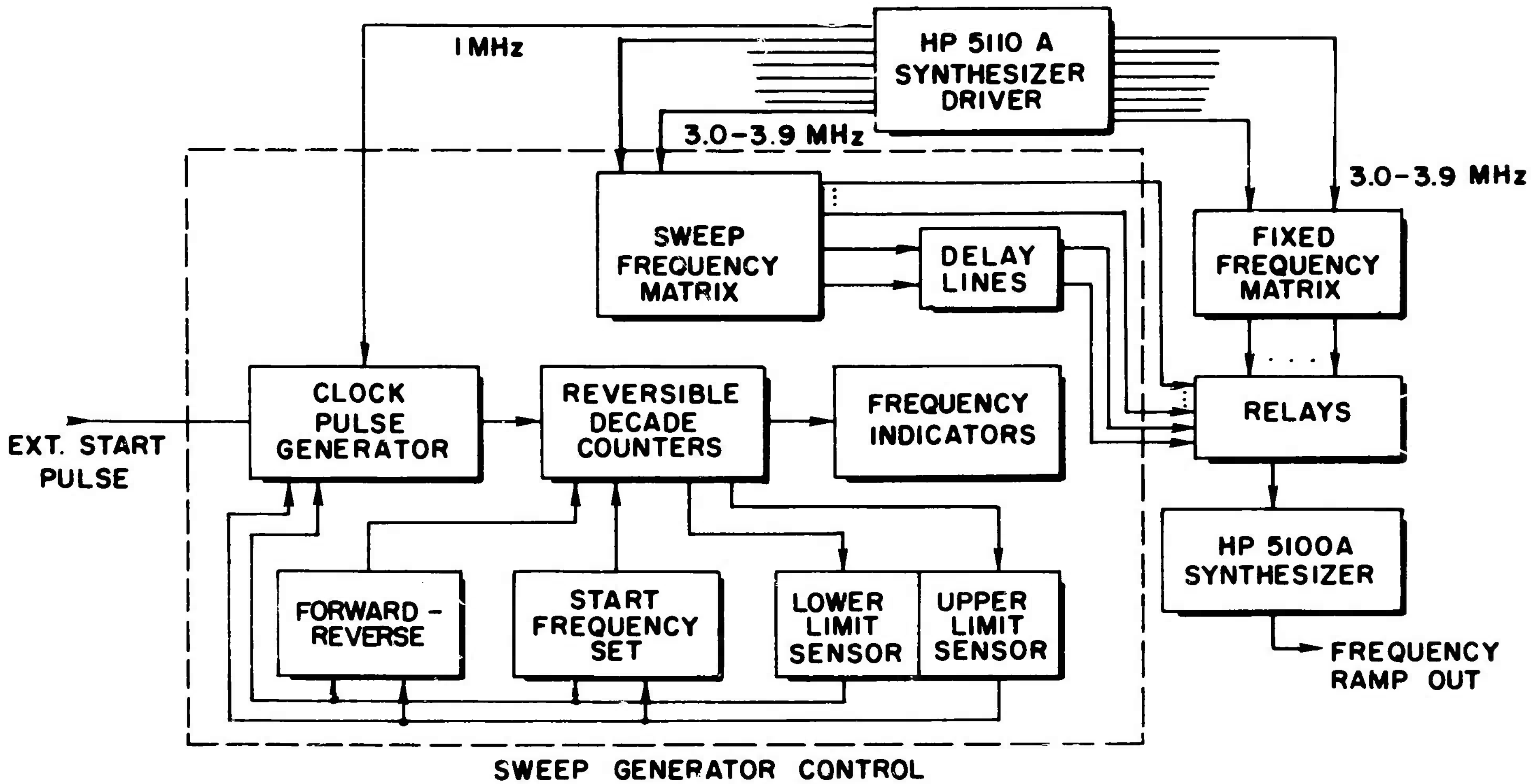


Fig. 3. BLOCK DIAGRAM, CHIRP II LINEAR-SWEEP GENERATOR.

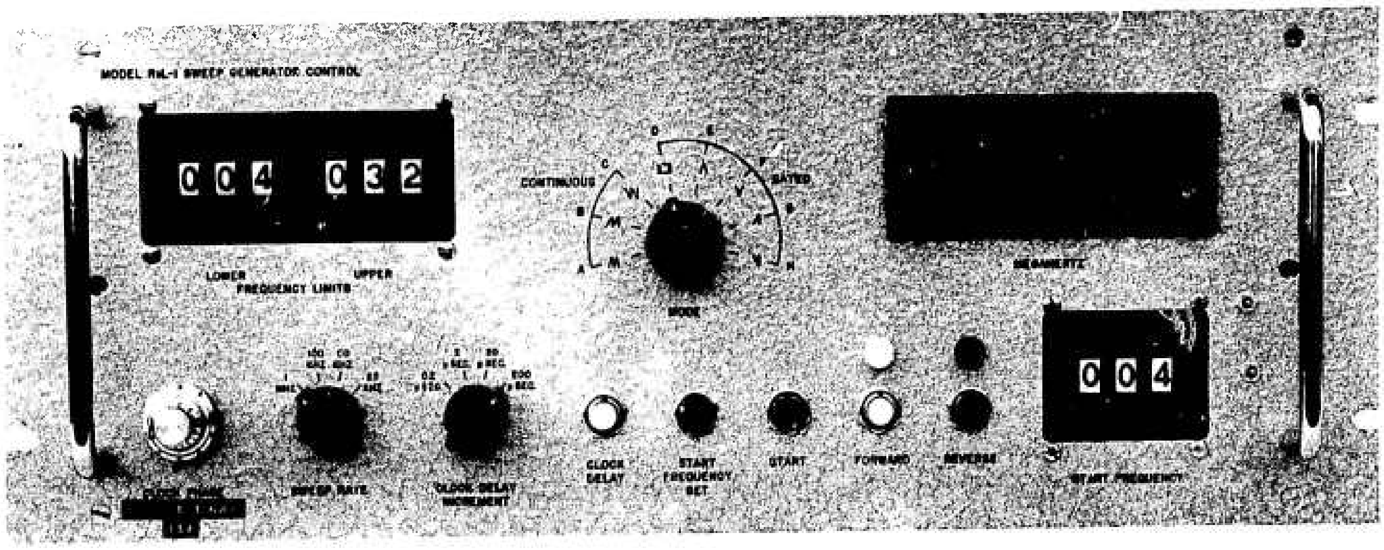


Fig. 4. FRONT PANEL, SWEEP GENERATOR CONTROL.

small p-p phase error, may be considered as the sum of the ideal linear phasor and an amplitude-modulated spurious signal in phase quadrature. The resulting sideband power has been calculated by Barry [Ref. 3] and is lower than the desired signal by an amount

$$11.5 + 20 \log \frac{1}{f_i t_i} \text{ dB} . \qquad (1)$$

For a sweep rate of 100 kHz/sec, frequency and time increments f_i and t_i are 1 Hz and 10 μsec, respectively. These values give a total sideband power of -111.4 dB relative to the ideal sweep signal. This finding implies that sidebands created by the frequency-stepping technique (if the step is small enough) should not in general be the limiting factor in determining resolution, and that other causes of PM and AM will set the sideband levels actually achieved. Some of these causes will be examined in the next section.

2. Sources of Spurious Components in Synthesized Linear-Sweep Generators

a. Phase Misalignment of Base Synthesizing Frequencies

Repetitive switching of base synthesizing frequencies which are not phase identical at the instant of switching will create a periodic phase modulation on the output frequency ramp. An ideal linear frequency sweep $v(t)$ can be represented as

$$v(t)_{ideal} = V_c \cos\left[\omega_o t + \frac{\alpha t^2}{2}\right] , \qquad (2)$$

where V_c = peak carrier voltage, ω_o = start frequency, and α = frequency sweep rate. The non-ideal case may be represented as

$$v(t)_{non-ideal} = V_c \cos\left[\omega_o t + \frac{\alpha t^2}{2} + (\Delta\phi)\ p(t)\right] , \qquad (3)$$

where $\Delta\phi$ = peak phase deviation of the synthesizing frequencies and $p(t)$ is a dimensionless function describing the phase error and having maximum and minimum values of +1 and -1, respectively. If ideal and non-ideal ramps having identical sweep rates are difference mixed, the resultant is a constant frequency which is phase-modulated, namely

$$v(t)_{difference} = V_d \cos\left[\omega_d t + (\Delta\phi)\ p(t)\right] , \qquad (4)$$

where ω_d is dependent on sweep start times and starting frequencies. This is the standard representation of a fixed-frequency, phase modulated wave.

Goldman [Ref. 9] has shown that a wave being phase modulated by a single frequency has an infinite number of sidebands in its frequency spectrum. These sidebands are separated from the carrier by the modulating frequency and its harmonics. When the modulation index is small, as in the present situation, the amplitude of the sidebands diminishes rapidly as the order of the harmonics increases.

If the wave is phase modulated by two or more frequency components, all cross-product sidebands (i.e., the sum and difference frequencies between the components and their harmonics) are generated. Again, the amplitudes diminish rapidly for the higher-order cross products when the modulation indexes are small.

The phase modulation function $(\Delta\phi)$ p(t) can be described as follows: The output of the switch matrix feeding any given decade is switched periodically to each of its ten input frequency lines in sucession. Ideally, the switchover between two successive frequencies should occur when their instantaneous phases are the same, so that no discontinuity appears in the output waveform. If no measures are taken to control the phases of the input frequencies (i.e., if an unmodified synthesizer is used) a phase discontinuity will appear in the output waveform at the time of each switchover, which may be of any value up to ±180 degrees. The phase difference between the resulting waveform and the ideal waveform with no discontinuities will be a function $(\Delta\phi)$ p(t) appearing somewhat like Fig. 5, having a period of ten switching intervals.

In the case where the synthesizer is being swept at 100 kHz/sec, analysis of the function for the 10^0 decade in the synthesizer shows that there is a fundamental modulation component (f_m) of 10 kHz, producing sidebands placed at 10 kHz on either side of the fixed difference frequency. Sidebands produced by harmonics of the 10-kHz modulation frequency are also present, being spaced from the carrier at integral multiples of 10 kHz. Likewise, synthesized frequencies in the 10^1 decade are being switched one-tenth as fast, i.e., at 10 kHz/sec, and the fundamental modulation frequency is 1 kKz, producing still another series of sidebands spaced at 1-kHz intervals from the fixed difference frequency. Decades 10^2 through 10^7 also contribute sideband spectrums spaced at 100-, 10-····0.001-Hz intervals, respectively. Additional sidebands resulting from the sum and difference frequencies of the six fundamentals and their harmonics make the frequency spectrum extremely complicated, though as noted above, the higher-order sideband components are of negligible amplitude, provided that the modulation index of the various fundamental modulation frequencies can be kept small.

Since the fixed frequency is the difference between ideal and non-ideal (staircase) sweeps, one can visualize that the above fixed sideband spectrum is, in effect, the sideband spectrum of the non-ideal staircase sweep; that is, the non-ideal sweep consists of a series of frequency ramps spaced about the fundamental ramp, whose amplitudes and

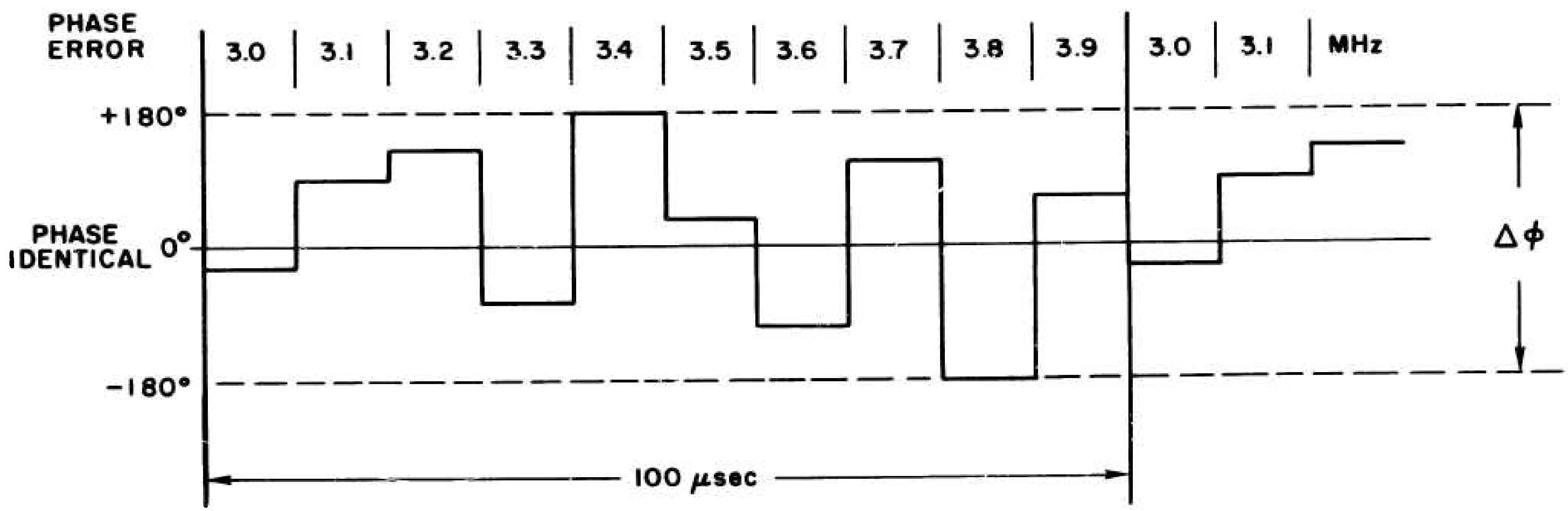


Fig. 5. TYPICAL PHASE DISCONTINUITY IN 3.0-3.9 MHz SYNTHESIZING FREQUENCIES IF NO PHASE ADJUSTMENTS ARE MADE TO BASE SYNTHESIZING FREQUENCIES. (Time scale shown for 10 degree decade at 100 kHz/sec sweep rate.)

intervals are determined by the phase errors between the successive 3.0- to 3.9- and 30- to 39-MHz synthesizing frequencies at switching times, and by the rates at which the frequencies are switched.

It becomes obvious that phase errors between base synthesizing frequencies at switching times must be reduced to a minimum to minimize sideband levels.

b. Non-Simultaneity of Waveforms in Mixer/Dividers

Signal propagation delays of several microseconds occur in each mixer/divider unit. These delays are principally determined by bandpass filter response characteristics. The delays cause the times of frequency switching in the various decades to be nonsimultaneous. Thus phase and frequency errors are generated at the synthesizer output at the time of each "carry."

The problem is illustrated by the following example: Consider a frequency change from 199,999 Hz to 200,000 Hz during a frequency sweep in which 1-Hz steps are taking place each 10 μsec. At this particular step, the 10^5 mixer/divider output changes upward 100,000 Hz, while the 10^4, 10^3, etc., mixer/dividers all change downward (90,000 Hz in the 10^4 decade, 9,000 Hz in the 10^3 decade, and so on down to 9 Hz in the 10^0 decade). Ideally, the combined effect of all these changes should appear at the last mixer simultaneously, so that only a 1-Hz change is seen at the output.

Now assume that the frequency switching occurs simultaneously in all decades, but that there is a time delay through each mixer/divider of 4 μsec. Figure 6 shows the signal path through the last three decades, and Fig. 7 shows the effect of these delays graphically. At the change from 199,999 to 200,000 Hz, the first effect that occurs at the last mixer in Fig. 6 is the upward change of 100,000 Hz at input C. For the next 4 μsec, then, the output frequency is 100,000 Hz too high, which causes the output phase to advance 144 degrees from the desired value. At 4 μsec, the drop of 90,000 Hz at input B reaches the output, so that the output frequency becomes 10,000 Hz high; therefore, the output phase gains another 14.4 degrees in the second 4-μsec interval. In the next three 4-μsec intervals, the phase gains by successively

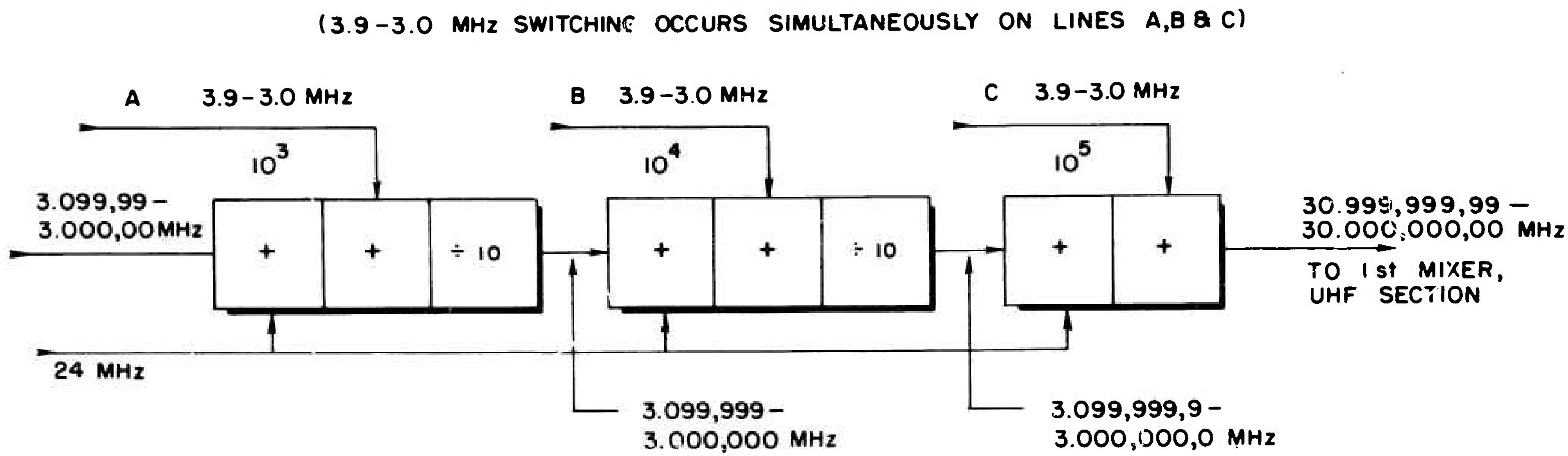


Fig. 6. BLOCK DIAGRAM OF 10^3 AND 10^4 DECADE MIXER/DIVIDERS AND 10^5 DECADE MIXERS.

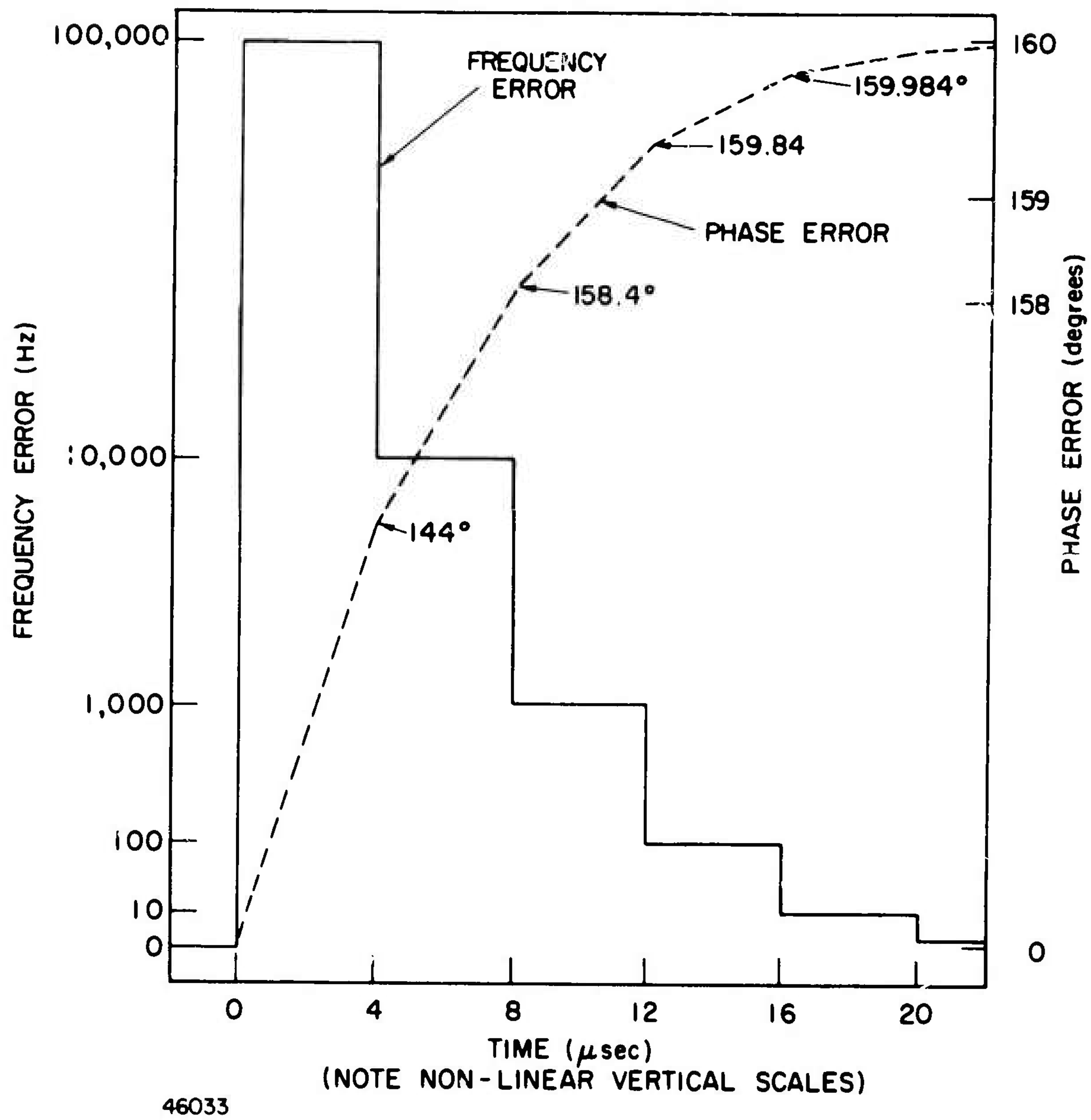


Fig. 7. CALCULATED FREQUENCY AND PHASE TRANSIENTS WHEN SWITCHING FROM 199,999 Hz to 200,000 Hz, ASSUMING 4 μsec DELAY PER DECADE.

smaller amounts. The total phase discontinuity approaches 160 degrees, of which over 158 degrees are attributable to the delays in the 10^4 and 10^5 decades.

The most obvious approach to reduction of sweep phase errors is reduction of time delays by broadening filter bandwidths. The delays can be significantly reduced by that means, but not to the extent that phase errors become tolerable.

c. Gate Noise

The diode, unbalanced gates supplied with the synthesizer, which repetitively switch the 3.0- to 3.9-MHz synthesizing frequencies, generate noise in the form of impulses at the switching rate. This noise can upset operation of the dividers in the mixer/divider units, causing step-function phase errors and, for the more significant decades, it can cause amplitude transients in the synthesizer output. Even when not serious enough to cause the divider circuit to lose count, the gating noise represents additive noise which will appear at the synthesizer output.

3. Reduction of Spurious Frequency Components

a. 3.0- to 3.9-MHz Phase Alignment

Chirp I 3-MHz comb phase alignment was limited to within ± 18 degrees--easily accomplished since each of the 3.0- to 3.9-MHz signals is derived in the 5110A fixed-frequency section by division from 30-39 MHz. Causing these divide-by-10 circuits to skip entire cycles at their inputs shifts the phase at their outputs in one-tenth-cycle increments.

More precise phase alignment of the 3.0- to 3.9-MHz synthesizing frequencies in Chirp II was accomplished by tuning of inductors L57 and L58 in mixer/divider units A5 through A14 in the 5110A Synthesizer Driver, after the phases were adjusted to the ± 18 degrees limitation described above. The 1 MHz output from the 5110A was used as a reference. The 3.0- to 3.9-MHz positive slope zero crossings were aligned to within an accuracy of ± 7 degrees from the positive-slope zero crossing of the 1-MHz reference.

Figure 8 illustrates the degree of phase alignment accuracy attained. The digital switching waveforms (used to switch among the synthesizing frequencies) were developed from the same 1-MHz reference. A variable 1-μsec delay line (the clock phase control on the front panel of the Sweep Generator Control unit, Fig. 2), inserted between the 1-MHz reference and the switching circuitry, provided for exact synchronization of the switching transition with the phase-aligned zero-crossings of the 3.0- to 3.9-MHz synthesizing frequencies (Fig. 8).

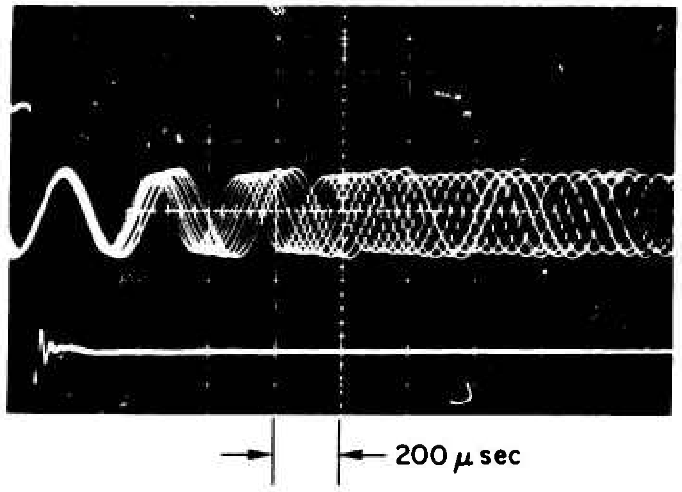


Fig. 8. 10-kHz SWITCHING WAVEFORM SUPERIMPOSED ON 10^3 DECADE SYNTHESIZING-FREQUENCY OUTPUT.

b. <u>30- to 39-MHz Phase Alignment</u>

Phase alignment of the 30- to 39-MHz synthesizing frequencies in Chirp I was not attempted. In Chirp II, in order to achieve output phase continuity over more than 1-MHz intervals, inductors L7 through L11 in mixer/divider units A5 through A14 in the 5110A Synthesizer Driver were adjusted to give phase-identical positive-going zero crossings of the 30-, 31-····39-MHz signals coincident with the same 1-MHz reference used for the 3.0- to 3.9-MHz frequencies. Adjustments were made to the 30- to 39-MHz frequencies before attempting alignment of the 3.0- to 3.9-MHz lines, since any phase adjustment of the higher frequencies affects phase of the lower ones.

Because these frequencies are high enough so that phase shift in measuring devices such as standard oscilloscopes, probes, etc., could introduce significant errors into the zero crossings over a range of 30-39 MHz, a Tektronix* 1S1 sampling plug-in, in conjunction with an RM 544 oscilloscope, was used for the initial alignment. After suitable alignment was attained, a regular 1A1 Tektronix dual-trace plug-in was substituted for the sampling unit, along with the standard probes to determine if there was enough measurement-device phase shift present to make future nonsampling alignment invalid. It was found that the 1A1

*Beaverton, Oregon

vertical plug-in with appropriate probes had negligible phase shift, and future phase alignment using this technique would indeed be valid. Phase alignment of the 30- to 39-MHz frequencies at critical switching times was accomplished to within an accuracy of ± 15 degrees at the positive-slope zero crossing of the 1-MHz reference.

c. Mixer-Divider Time Delay

Time delays in the last half of the 10^3 decade mixer/divider unit, the entire 10^4 decade mixer/divider unit, and the 10^5 decade mixers were reduced by changing the bandpass characteristics of the filters located between the mixers. Obviously, the time delays cannot be reduced indefinitely by this means, since one has to accept a compromise between time delay and rejection of unwanted mixing products. Figure 9 shows time delays both before and after filter modification. Time delays between base synthesizing-frequency injection points were reduced from 3.5 to 1.8 μsec. The expected total output phase error for 100 kHz carry operations at a sweep rate of 100 kHz/sec was thus reduced from 160 degrees to approximately 80 degrees.

d. Delay Lines

To further reduce the phase error from the 80 degrees obtained by reducing time delay as described above, residual time delay in the mixer/dividers was compensated by a delay-line scheme.

If a time delay equal to the time it takes for the synthesized signal to travel between mixing points in the 10^3 and 10^4 decades is inserted in the 10^4 decade base synthesizing-frequency line, then transitions in the synthesizing and synthesized signals arrive simultaneously at the mixer in the 10^4 decade. Likewise, if a larger time delay, equal to the time for the synthesized signal to travel between mixing points in the 10^3 and 10^5 decades, is inserted in the 10^5 decade base synthesizing-frequency line, again the synthesizing and synthesized signal transitions will arrive at the last mixer simultaneously. Figure 10 shows placement of the delay lines in the 10^4 and 10^5 decade base synthesizing-frequency lines.

Delay lines with delays of 1.8 μsec (equal to residual delay in one mixer/divider) and 3.6 μsec (the delay in two cascaded

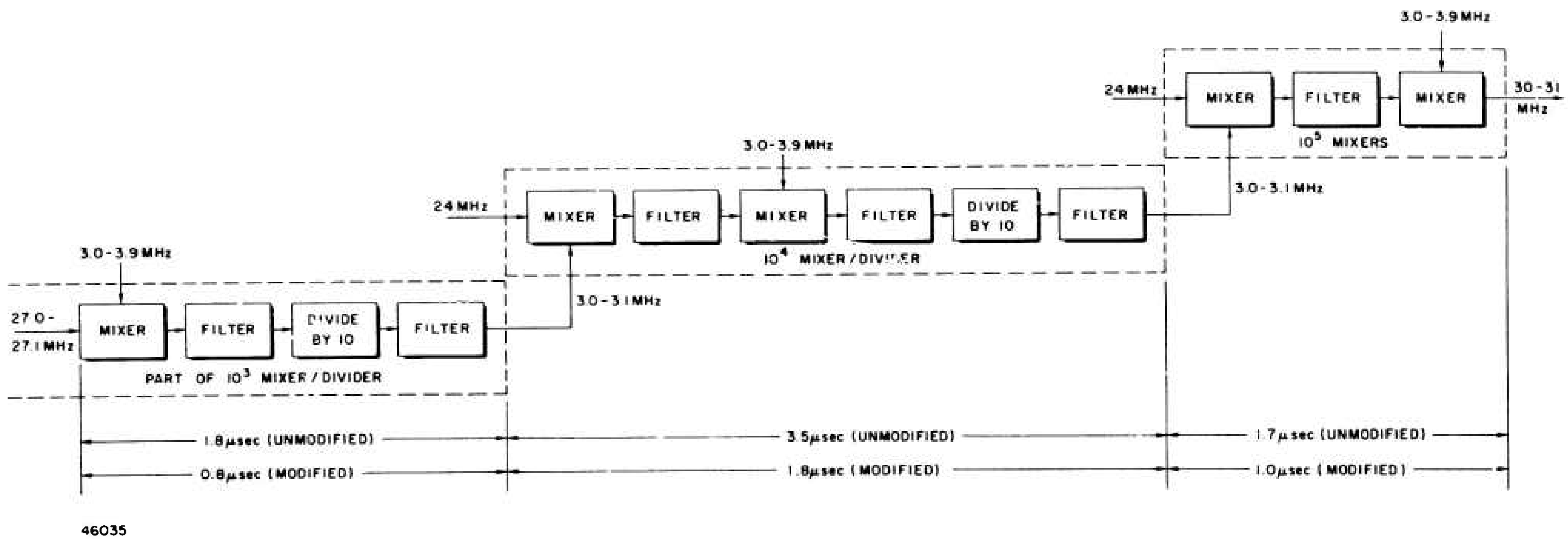


Fig. 9. SIMPLIFIED BLOCK DIAGRAM, 10^3, 10^4 AND 10^5 DECADES, SHOWING DECREASE IN TIME DELAY OBTAINED BY FILTER MODIFICATION.

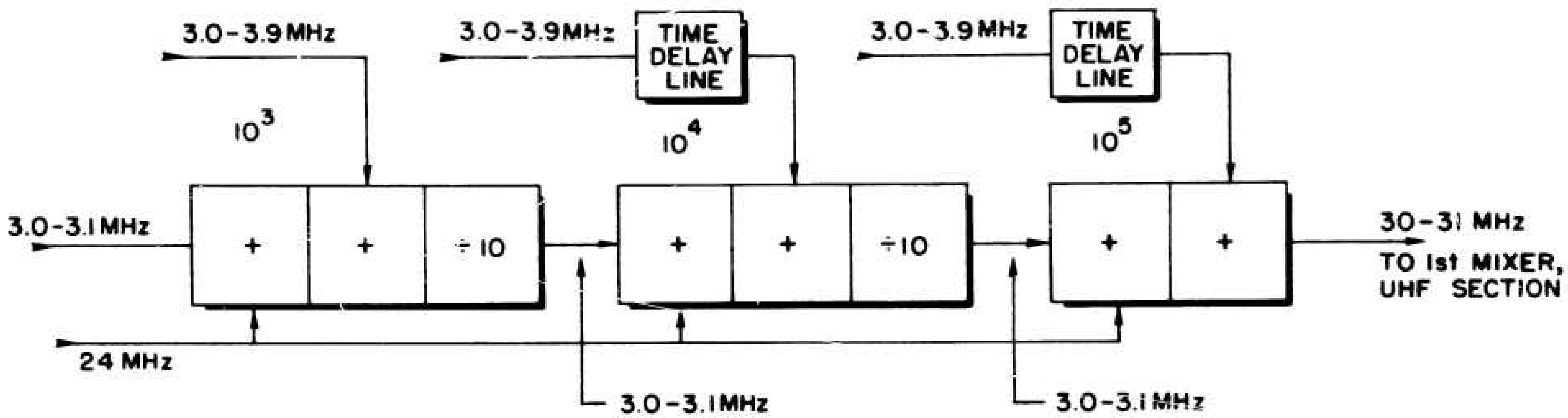


Fig. 10. BLOCK DIAGRAM OF 10^3 AND 10^4 DECADE MIXER/DIVIDERS AND 10^5 MIXERS, SHOWING PLACEMENT OF DELAY NETWORKS IN SYNTHESIZING FREQUENCY LINES.

mixer/dividers) were ordered for installation in the 10^4 and 10^5 decade base synthesizing-frequency lines, but they did not arrive in time for installation prior to making the system measurements to be described in Chapter IV. Instead, a 1000-ft length of RG-8 coaxial cable was used as a temporary delay line in the 10^5 decade base synthesizing-frequency line, providing a delay of approximately 1.5 μsec. The 10^4 decade lines had no delay inserted during the system measurements and initial soundings described in this report.

(It would seem that the delay line scheme could, in principle, remove the requirement for broadening the mixer/divider filter bandwidths, provided that the effect of a filter could be characterized adequately by a single time delay. However, the effective time delay varies across the 3- to 4-MHz bandwidth especially near the skirts of the filter, and hence it is desirable to use both filter-broadening and delay compensation.)

e. Balanced Gates

To reduce switching-rate impulse noise arriving at the mixer/divider via synthesizing frequency lines, diode gates in the 3- to 4-MHz switch matrix were replaced by balanced versions which consisted of two toroid transformers and two fast-switching diodes. Small printed circuit boards of the same dimensions as the units bearing the original unbalanced gates, but supporting balanced gate circuitry, were installed in the switch matrix casting. Figure 11 shows the matrix casting, along with one of the old and one of the new gate boards. Residual transients on the synthesizing frequency signal at the critical switching times are negligible, as can be seen in Fig. 8.

C. Switching Between Sweep- and Fixed-Frequency Operation

Digital switching waveforms are applied to the 3- to 4-MHz switch in the standard 5100A synthesizer via three 50-pin connectors on the rear panel. The conductors from the connectors are then bundled together and routed to the pushbutton switch on the front panel. From this point, conductors are fed to the 3- to 4-MHz switch. Total length of the cable from rear panel to the 3- to 4-MHz switch matrix via this

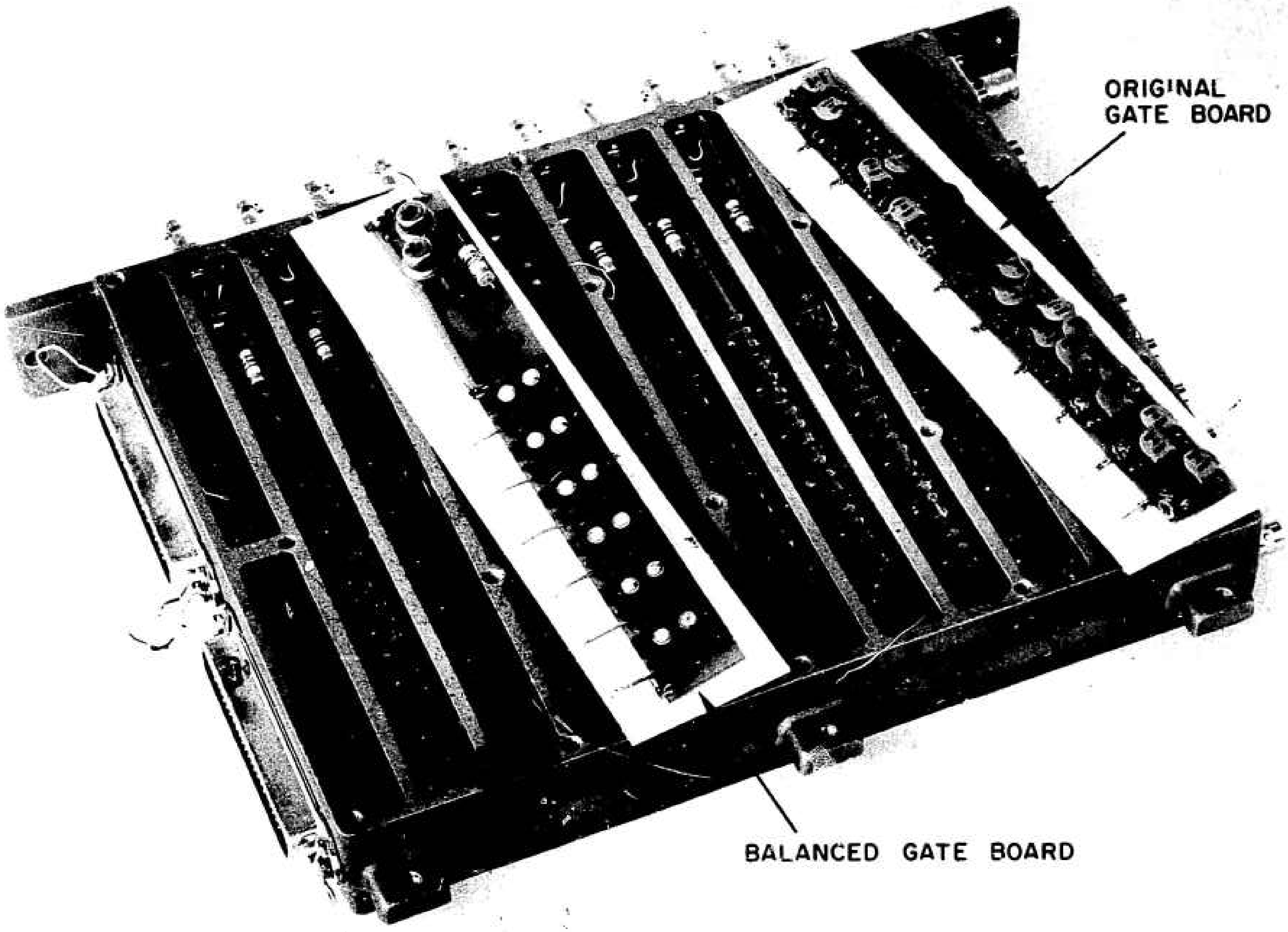


Fig. 11. 3- TO 4-MHz SWITCH MATRIX, WITH OLD AND NEW (BALANCED) GATE BOARDS.

route is approximately 3 ft. The switching waveforms, having rise and fall times on the order of 15-40 nsec, can be degraded by the capacitance introduced in such a long run of bundled cables. It was thought that if the 3- to 4-MHz switch matrix was relocated to be near the digital circuitry (contained in the Sweep Generator Control unit), degradation of switching waveforms could be reduced significantly. However, if the switch matrix was removed from the actual synthesizer, use of the synthesizer as a fixed-frequency device controlled by the pushbuttons on the front panel would be impractical.

This problem was solved by adding a second switch matrix, to be used only for fixed-frequency operation. This unit was located in the normal position in the synthesizer. To switch from fixed- to sweep-frequency mode, a set of single-pole, double-throw, coaxial relays in the lines to the mixer/dividers was used to choose signals from either matrix. The normally closed position of the relays was chosen for the sweeping mode, since the synthesizer would ordinarily be used in this condition most of the time. To switch to fixed-frequency operation, it was necessary only to apply power to the relay coils (accomplished by switching the local-remote switch on the front of the synthesizer).

To make room for the relays in the synthesizer package, the mixer/dividers that generate the two least significant figures (10^{-1} and 10^{-2}) were removed.

With the digitally swept switch matrix mounted in the Sweep Generator Control near the digital circuitry, switching waveforms applied to the balanced gates were not degraded by long cable runs. The six 3.0- to 3.9-MHz outputs from the matrix were connected to the relays in the synthesizer via coaxial cables. Figure 12 shows the location of the swept switch matrix in the Sweep Generator Control.

D. Sweep Formats

Projected applications for Chirp II dictated the availability of sweep formats with positive and negative slopes. When used in a radar, positively and negatively sloped sweeps could be used to resolve the ambiguity between range and doppler. For making high-power backscatter

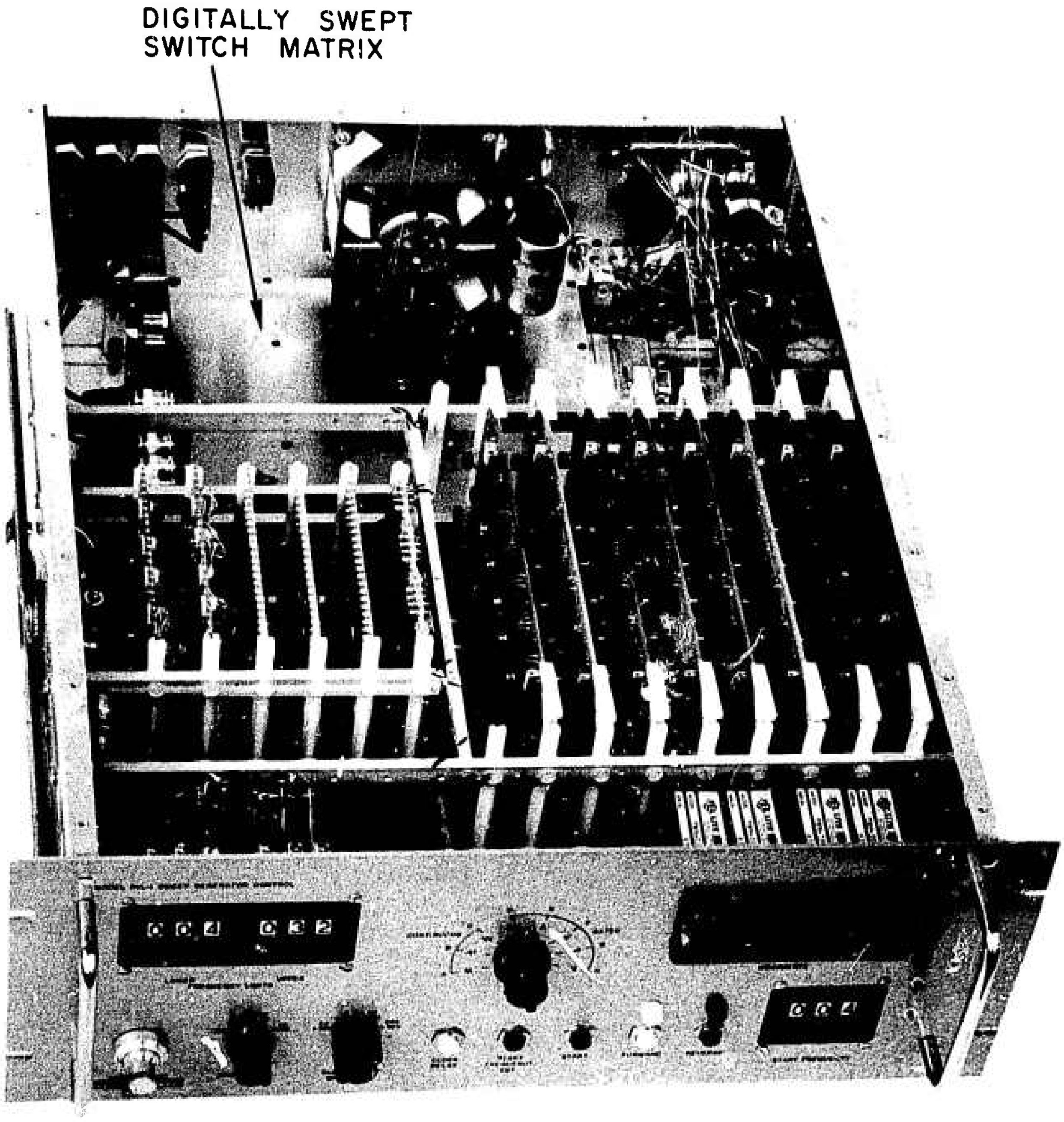


Fig. 12. SWEEP GENERATOR CONTROL, SHOWING DIGITAL CIRCUITRY AND DIGITALLY SWEPT SWITCH MATRIX.

measurements, the availability of the positively and negatively sloped sweeps was necessary, since use of a frequency-tracking, servo-tuned power amplifier required frequency sweeps with no abrupt frequency discontinuities.

Sweep formats in Chirp II are divided into two basic categories: (1) continuous "sawtooth" or "triangular" sweeping and (2) gated sweeps, initiated by "start" pulses.

The eight available formats are listed in Table 2. Figure 4 shows the format selector (MODE), manual FORWARD and REVERSE selectors, CLOCK DELAY control, and other manual controls.

Gated or continuous sweeping can be initiated by an externally generated timing pulse, or manually, by front-panel pushbutton. Provision is made to delay the ramp in time by intervals of 0.2, 2.0, 20 and 200 msec, and thereby allow the difference frequency between the transmitted and demodulating ramps to be accurately adjusted.

In gated sweeps, either the upper or lower frequency limit terminates the sweep, depending upon the format used (i.e., "flyback" at the end of a sweep is optional).

Table 2

SWEEP FORMATS -- CHIRP II

I. Continuous

A.* Triangular sweep, both positive and negative slope.

B. Sawtooth sweep, positive slope.

C. Sawtooth sweep, negative slope.

II. Gated

D. Alternate positively and negatively sloped sweeps, each sweep initiated by a start pulse and terminated by the appropriate frequency limit.

E. Single triangular sweep beginning with positive slope.

F. Single triangular sweep beginning with negative slope.

G. Sawtooth sweep, positive slope.

H. Sawtooth sweep, negative slope.

*Letters correspond to lettered positions on MODE switch, Sweep Generator Control, Fig. 4.

E. Digital Circuitry

1. Decade Counters

Digital switching circuits used to determine the synthesizing frequencies for ramp generation were constructed of discrete components in Chirp I. Each synthesizer decade had an associated ring counter operating in synchronism with counters in the remaining decades. The ring counter was a unidirectional device; therefore, it was unsuitable for application in Chirp II, where a bidirectional counter was necessary for generating positively and negatively sloped ramps.

A bidirectional, modulo-10 shift counter constructed with Fairchild RTμL industrial micrologic integrated circuits was the basic digital switching block used in Chirp II. The counter, along with its associated synchronous carry logic and gate-driving circuitry, was constructed on one double-sided etched circuit board, as shown in Fig. 13. There are eight counters in each sweep generator control. Six are associated with the 3.0- to 3.9-MHz switch matrix, providing switching waveforms for synthesis of significant digits in the 10^0 through 10^5 columns. The two remaining counters supply switching waveforms to the 30- to 39-MHz and 350- to 390-MHz switches in the synthesizer, to determine significant digits in the 10^6 and 10^7 columns.

2. Clock-Pulse Generation

Clock-pulse generation is accomplished by division of the 1-MHz standard frequency obtained from the HP 5110-A Synthesizer Driver. The 1-MHz sinusoid is squared and divided to yield 100-kHz, 50-kHz, and 25-kHz outputs, which provide the lower three sweep rates. When sweeping at the 1-MHz/sec rate, the 10^0 decade shift counter is disabled, and the 100-kHz signal is fed to the 10^1 decade counter. Thus the synthesizer output frequency is switched in 10-Hz steps at the 1-MHz/sec rate. Selection of sweep rate is accomplished with the four-position SWEEP RATE rotary switch on the Sweep Generator Control front panel. Switching of clock pulses from 10^0 to 10^1 decade when using the 1-MHz/sec sweep rate is performed in conjunction with clock frequency selection, by the same front-panel control.

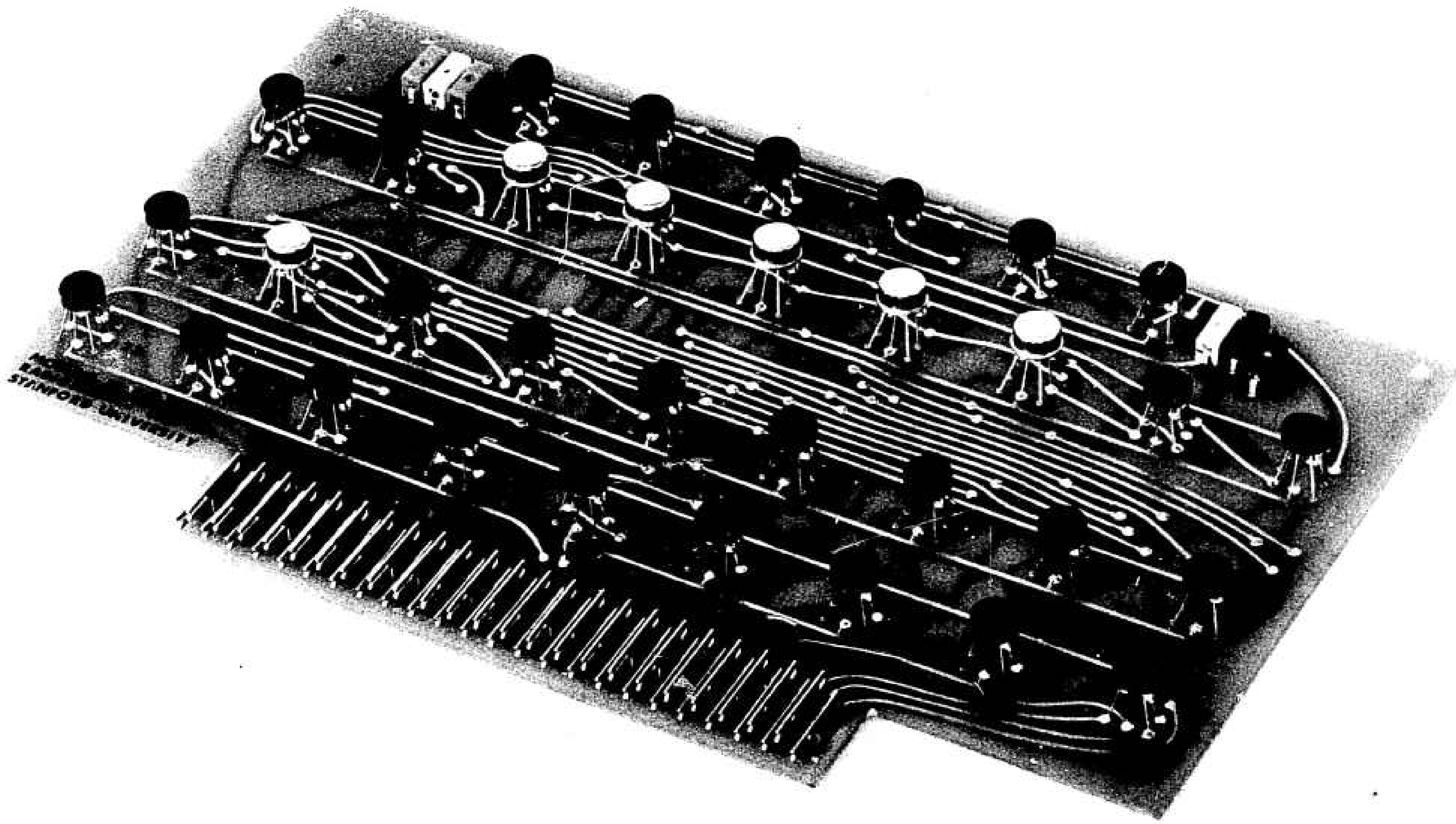


Fig. 13. REVERSIBLE DECADE COUNTER WITH ASSOCIATED GATE DRIVERS AND SYNCHRONOUS CARRY LOGIC.

3. Control Circuits

a. Limit Sensors

Frequency-sweep limit sensing is determined by digits in the 10^5, 10^6, and 10^7 Hz decades. A sensing signal from one of the ten digits in each decade is chosen by one of three single-pole, ten-position thumb-wheel switches and routed to a three-input "and" gate. When the selected frequency is reached, signals are present at all three inputs of the "and" gate, which triggers a one-shot multivibrator. This either reverses the decade counters or resets them to a starting frequency, depending on the setting of the MODE switch.

Both UPPER and LOWER FREQUENCY LIMIT sensing switches are incorporated in Chirp II and are shown in Fig. 4.

b. Establishing Initial Frequency

While the limit switches determine an upper and lower limit at which the frequency ramp may be terminated, they are unable to establish an initial count in the decade counters corresponding to some predetermined output frequency. The modulo-10 shift counters contain provision for setting any desired digit by establishing initial conditions in five JK flip-flops. Ten lines (two to each flip-flop) from each of the upper three decades are attached to the START FREQUENCY SET thumb-wheel switches (Fig. 4). A pushbutton (START FREQUENCY SET) on the generator control panel resets the 10^5, 10^6 and 10^7 decades to the selected frequency; all lower decades are reset to zero.

c. Frequency Indication

Four neon decade frequency MEGAHERTZ indicators (Fig. 4) mounted on the Sweep Generator Control panel indicate frequency to the nearest hundredth of a megahertz.

IV EQUIPMENT PERFORMANCE

Tests were conducted to measure attained sidelobe levels. The following paragraph gives a description of the measurement technique used, and Fig. 14 is a block diagram of the test setup.

Two linear sweep generators were programmed to sweep with a positive sawtooth slope at 100 kHz/sec over the 4- to 30-MHz frequency range. The outputs of both synthesizers were fed into balanced mixer "A." Sweeps were initiated at times separated by 1.2 sec, giving a difference frequency of 120 kHz at mixer "A" output. The 120-kHz signal was then fed into mixer "B" with a second frequency selected so that the output of mixer "B" was approximately 2 Hz. This signal was recorded on a magnetic-tape FM recorder at 1-7/8 IPS and on a paper chart recorder.

The purpose of the 1.2-sec offset of start times of the two frequency sweeps was to prevent coincident occurrence and possible cancellation of the phase discontinuities associated with 100-kHz and 1-MHz switching intervals.

As previously noted, the compensating delay lines for insertion in the base synthesizing-frequency lines in the 10^5 and 10^4 decades had not arrived at the time of these tests. The measurements were made with a coaxial line of 1.5-μsec delay inserted in the 10^5 decade line, and no delay in the 10^4 decade line. As a result, step-function phase jumps caused by time delays in the 10^3-10^5 mixer dividers were not compensated. A delay of 1.5 μsec in the 10^5 decade reduces the 65 degree step function phase error at 100 kHz transitions (10^5 decade only) to 10.8 degrees. Combined with approximately 6.5 degrees of phase error from the 10^4 decade mixer/divider and the 2 degree error from the remaining mixer/dividers, a phase jump not greater than 20 degrees was expected at 100-kHz transitions.

Figures 15a and 15b show several cycles of mixer B output recorded on the paper chart recorder. Transitions at 17.7 and 17.8 MHz for sweep generator 1 (Fig. 15a) show very small phase jumps. However, at the 17.9 MHz transition, a step function phase jump of approximately 33 degrees appears. For generator 2, a similar situation exists, showing very small phase jumps at 17.6 and 17.7 MHz, but a phase jump of approximately 30 degrees at the 17.8-MHz transition. Both of the large phase jumps

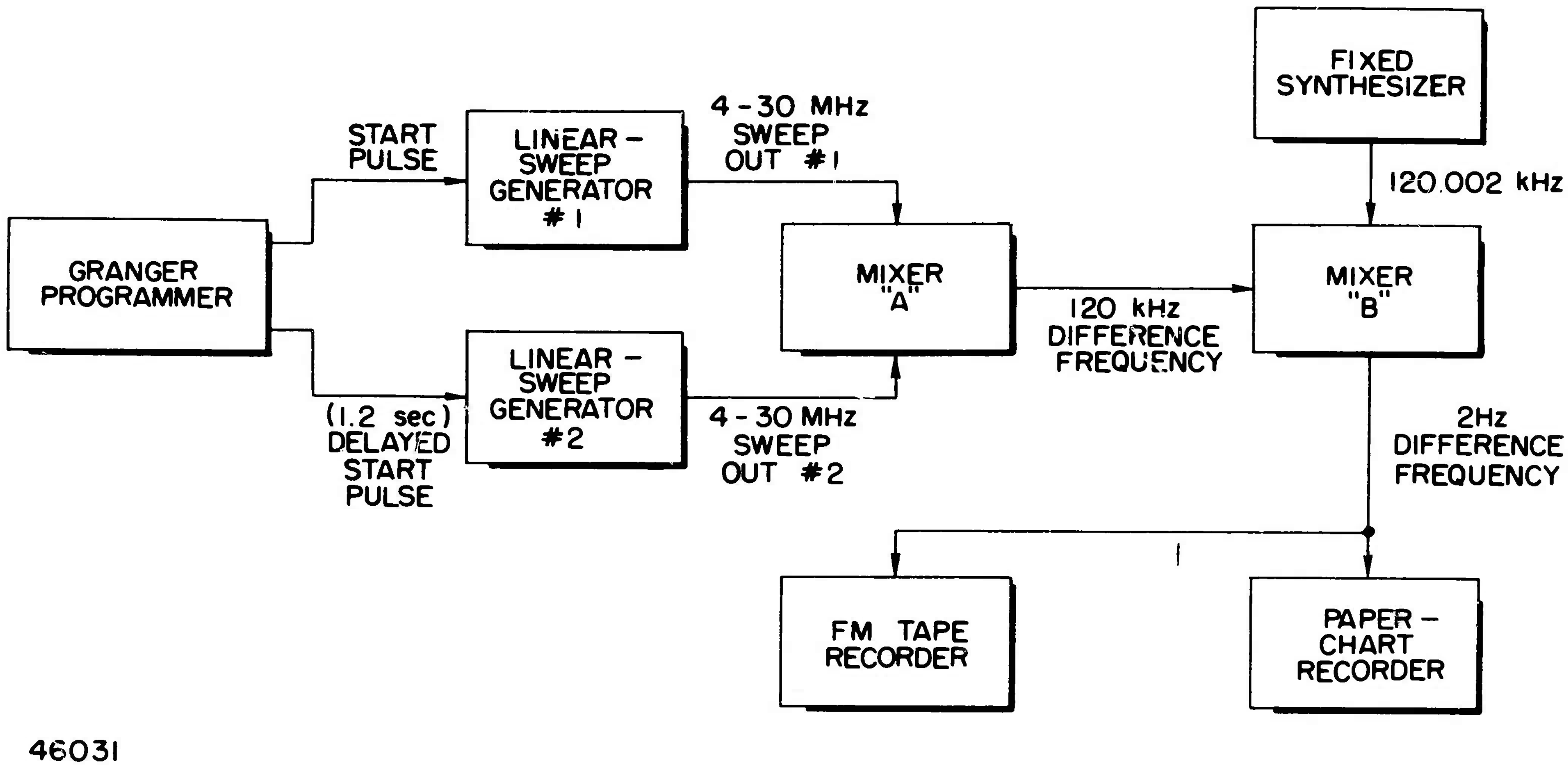


Fig. 14. BLOCK DIAGRAM OF TEST SETUP FOR MEASURING PHASE DISCONTINUITIES IN LINEAR SWEEP GENERATOR OUTPUT.

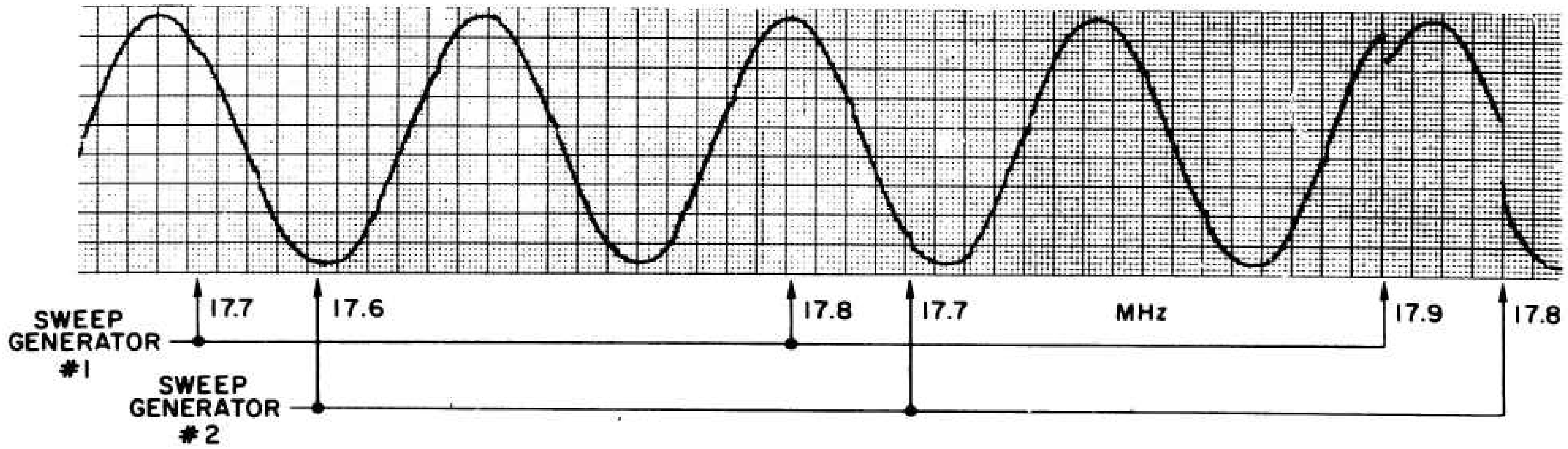


Fig. 15a. 2-Hz DIFFERENCE-FREQUENCY OUTPUT FROM MIXER B (FIG. 14) SHOWING PHASE JUMPS AT 100-kHz INTERVAL TRANSITIONS.

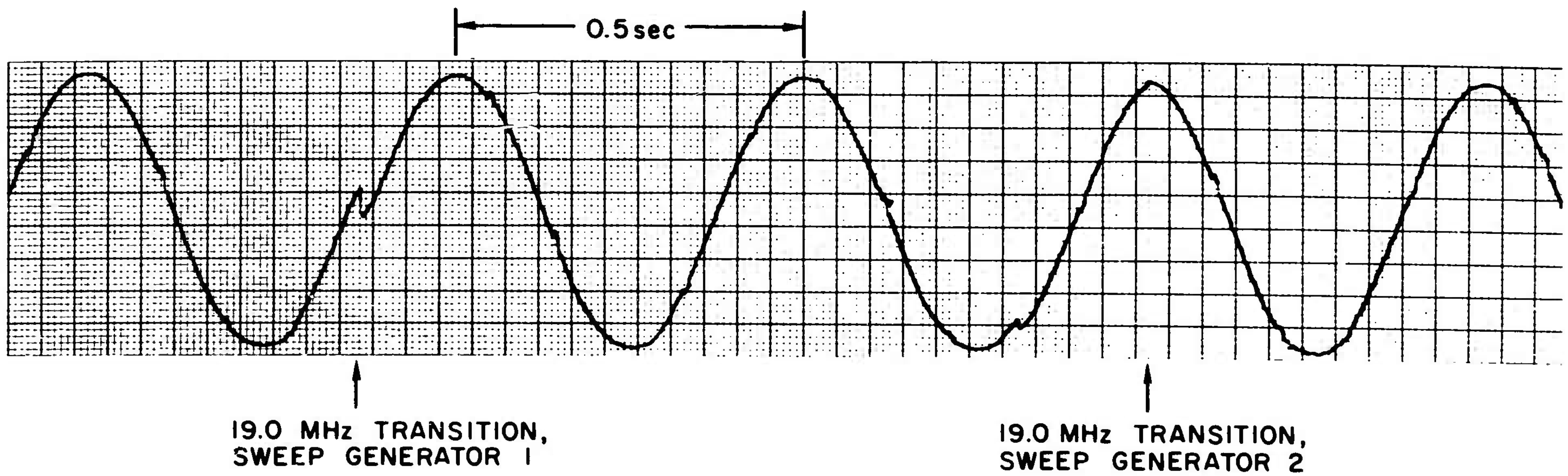


Fig. 15b. 2-Hz DIFFERENCE-FREQUENCY OUTPUT FROM MIXER B (FIG. 14) SHOWING PHASE JUMPS AT 1-MHz TRANSITIONS.

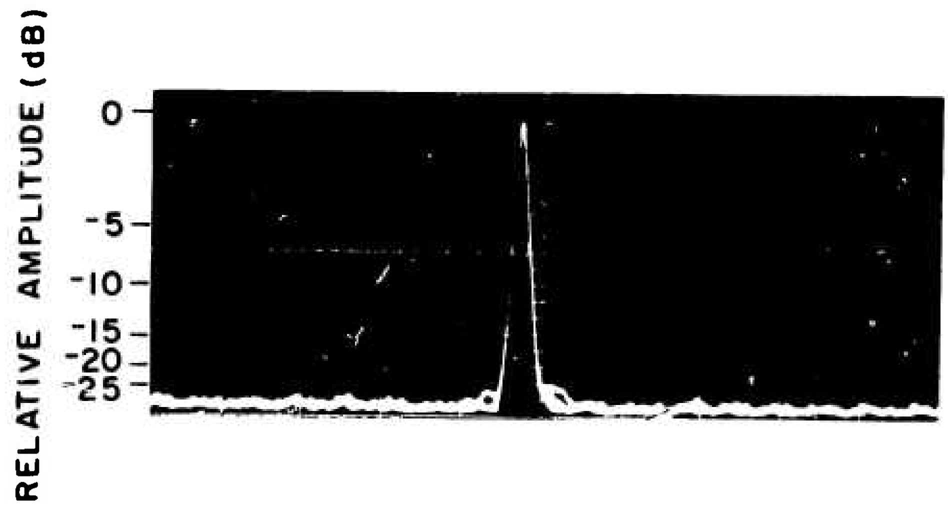
exceed the 20-degree maximum value calculated for compensation with the 1.5-μsec delay line by approximately 10 degrees. However, assuming a ± 10 degree tolerance on phasing of the 3-MHz comb frequencies, jumps in the range of 10-30 degrees should be expected. Measured phase jumps appeared to be confined in this approximate range.

In Fig. 15b, phase jumps occurring at the 19.0-MHz transition are shown; these were approximately 25 degrees for both sweep generators. Most megahertz-transition phase errors observed were typically 25 degrees, with the worst case approximately 80 degrees.

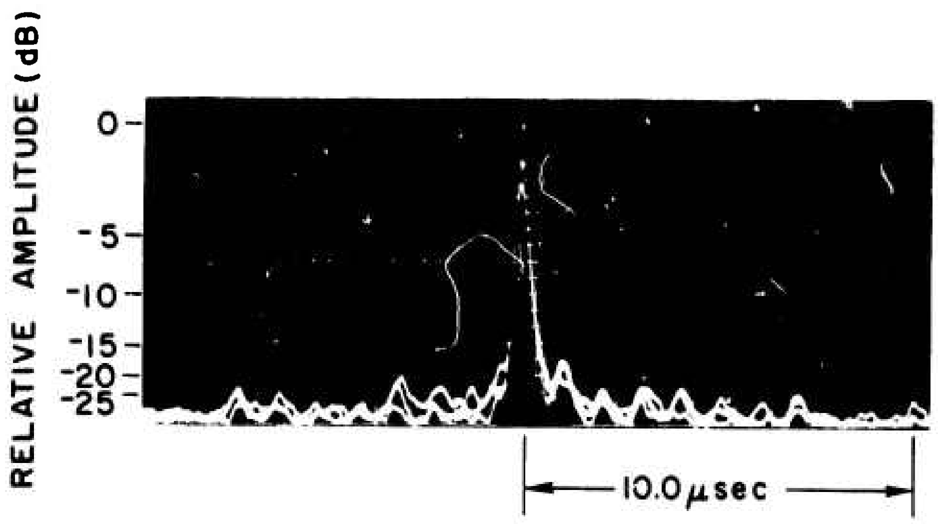
The 2-Hz data recorded on magnetic tape at 1-7/8 IPS was sped up by a factor of 1024 and recorded again to give an approximate frequency of 2 kHz for spectrum analysis by a 4-kHz bandwidth analyzer. Speed-up of the data by a factor of 10^3 makes it possible to determine the 3-dB width of the principal sweep-frequency component and to inspect the sidelobe spectra within the equivalent of a few microseconds of the principal sweep.

Figure 16a shows the spectrum of a 2-Hz audio oscillator signal recorded on the tape, sped up and re-recorded, and spectrum-analyzed as described above to show the capability of the tape equipment and analyzer alone. Figure 16b shows a spectrum analysis of the 2-Hz difference-frequency output from the test setup of Fig. 14. The principal sweep component appears with sidelobe spectra out to ± 10 μsec. The 3-dB width of the principal component is 50 nsec, while sidelobe levels are approximately 20 dB down.

Figures 17a and 17b are similar to Figs. 16a and 16b, except that the horizontal scale has been expanded to show detail close to the fundamental sweep component.

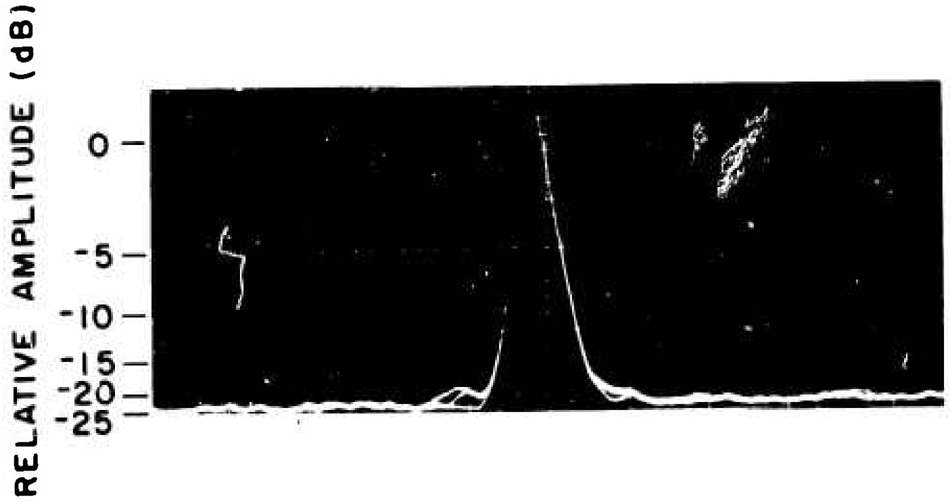


(a) Response to clean 2-Hz signal

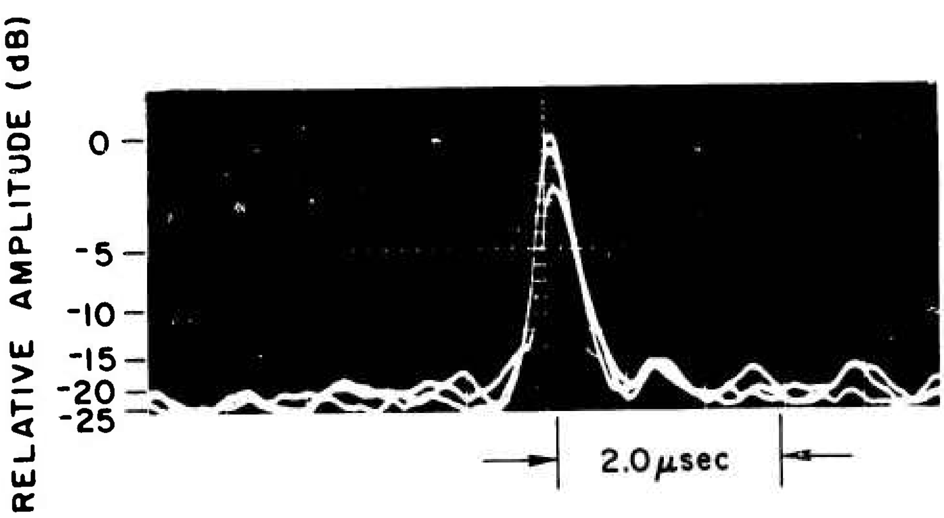


(b) Response to frequency-sweep generator difference-frequency output

Fig. 16. SPECTRUM ANALYSIS OF FREQUENCY-SWEEP OUTPUT (2-Hz SIGNAL FROM MIXER B, FIG. 14, RE-RECORDED AT HIGH TAPE SPEEDUP FACTOR).



(a) Response to clean 2-Hz signal



(b) Response to frequency-sweep generator difference frequency output

Fig. 17. SPECTRUM ANALYSIS OF FREQUENCY-SWEEP OUTPUT (2-Hz SIGNAL FROM MIXER B, FIG. 14 , SAME AS FIG. 16 EXCEPT HORIZONTAL SCALE EXPANDED.

V OBLIQUE SOUNDING TESTS

A. Equipment

First soundings using Chirp II sweep generators were conducted in December 1966 over an 1880-km path. The transmitting equipment was located at Texas Technological College, Lubbock, Texas. Receiving equipment was installed at Stanford. The frequency range transmitted was 4-30 MHz, and a power level of 100 W was used. Figures 18a and 18b are block diagrams of the the transmitting and receiving equipment; Fig. 19 is a photograph of the receiving equipment installation.

The I-F strip used was a Hallicrafters-Manson MHR-2S receiver. This receiver was well suited for the task because (1) all mixer injection signals are derived from a single frequency standard; (2) one available final I-F bandwidth is 6 kHz; this bandwidth permits a 3-msec time delay "window" at a 1-MHz/sec sweep rate; (3) long-time-constant AGC control is available; and (4) a product detector output is available--this output had a frequency range compatible with the available spectrum analyzer.

Two methods of recording ionograms were used. In the first, real time method, the nominal 100-Hz output of the receiver was spectrum-analyzed and presented on the monitor oscilloscope, whose horizontal scan was triggered by a start pulse from the programmer. The vertical sawtooth and video modulation were furnished by the spectrum analyzer to produce an ionogram in the usual format. This could be directly viewed or photographed by a Polaroid camera. Figure 20 shows an ionogram of this type.

The second method, while not operating in real time, allowed the received signal to be processed to a wide range of time-delay resolutions. The receiver output signal was recorded on tape and then played back at a speedup factor of from 2 to 32 times into the spectrum analyzer. The spectrum analyzer sawtooth output was connected to provide the horizontal scan to a Tektronix 543-B oscilloscope. The trace was recorded on 35 mm film moved past the lens of a Grass Kymograph camera. The video signal from the spectrum analyzer was connected to the oscilloscope Z-axis input. The video signal could also be connected to the

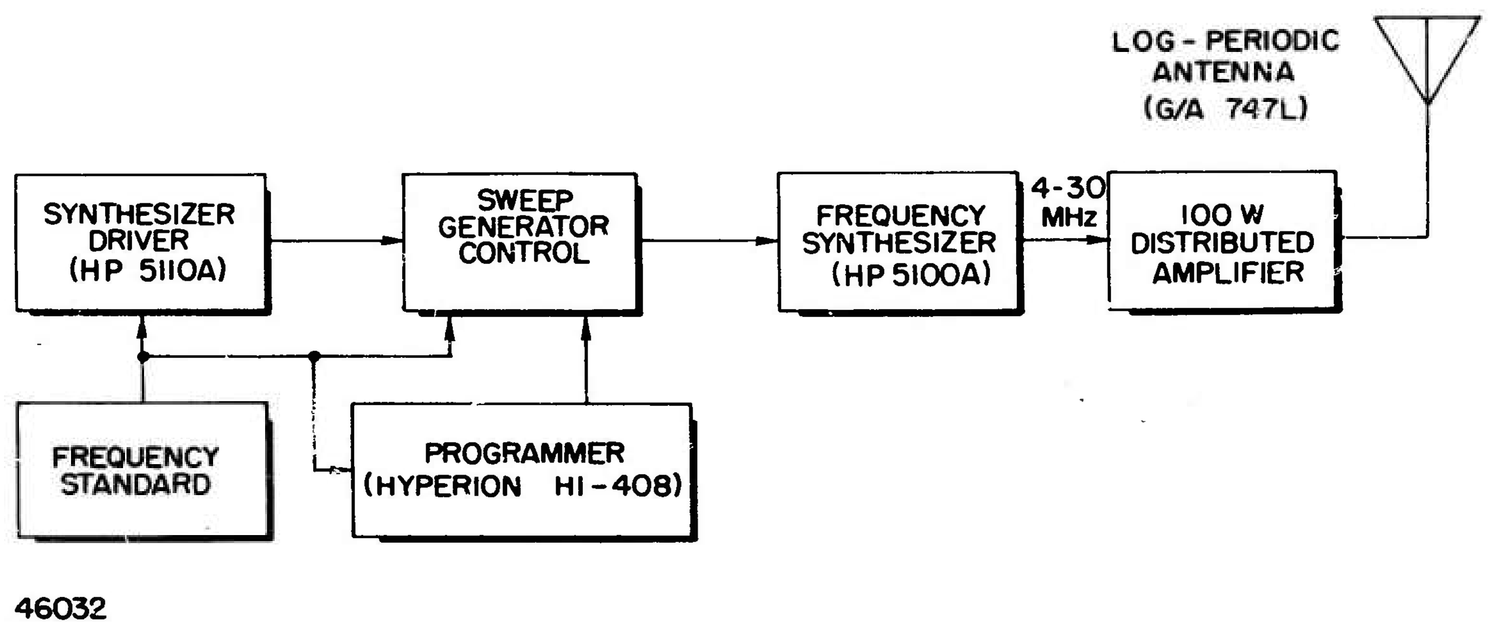


Fig. 18a. BLOCK DIAGRAM, CHIRP II OBLIQUE-INCIDENCE TRANSMITTING EQUIPMENT.

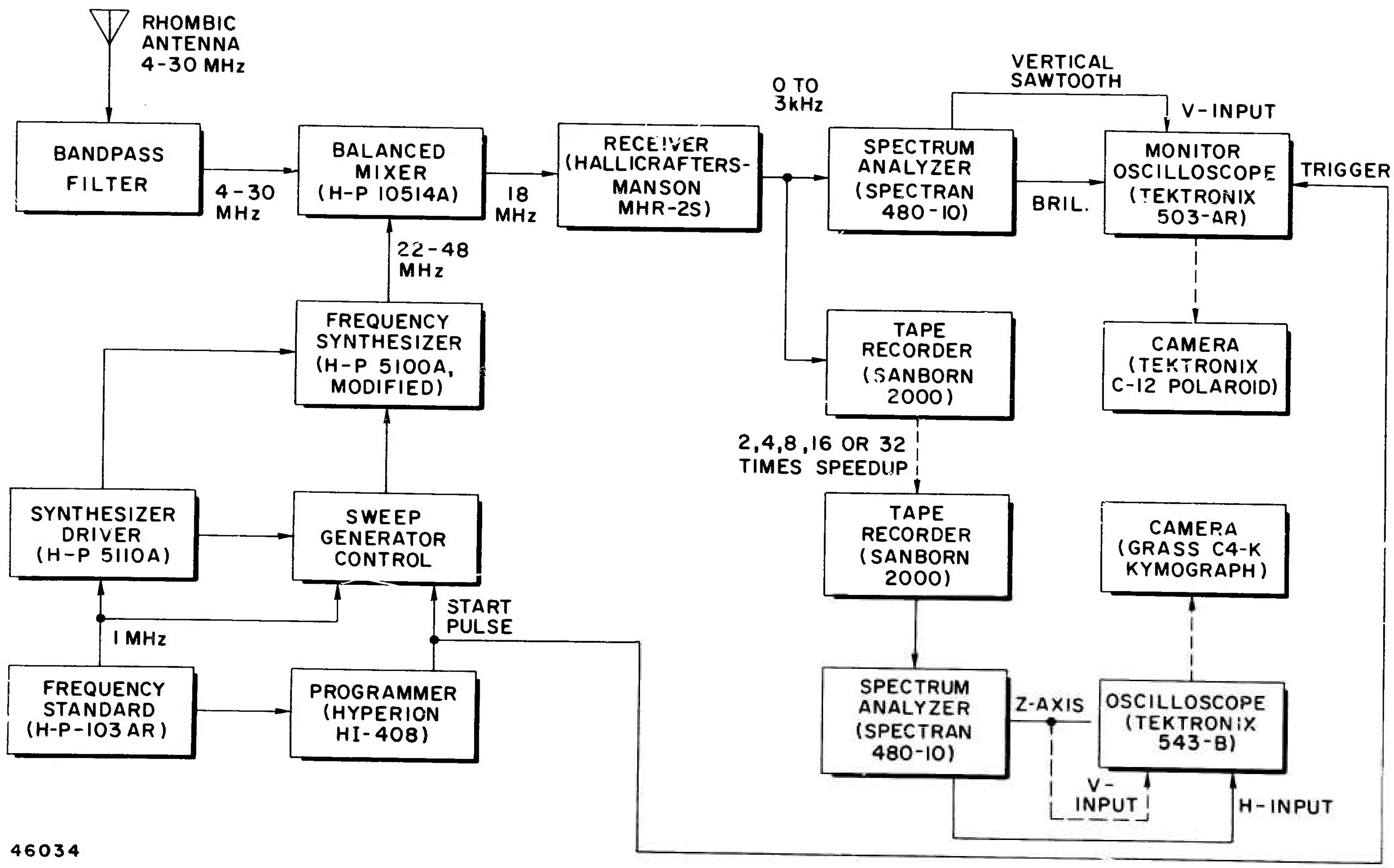


Fig. 18b. BLOCK DIAGRAM, CHIRP II OBLIQUE-INCIDENCE SOUNDER RECEIVER.

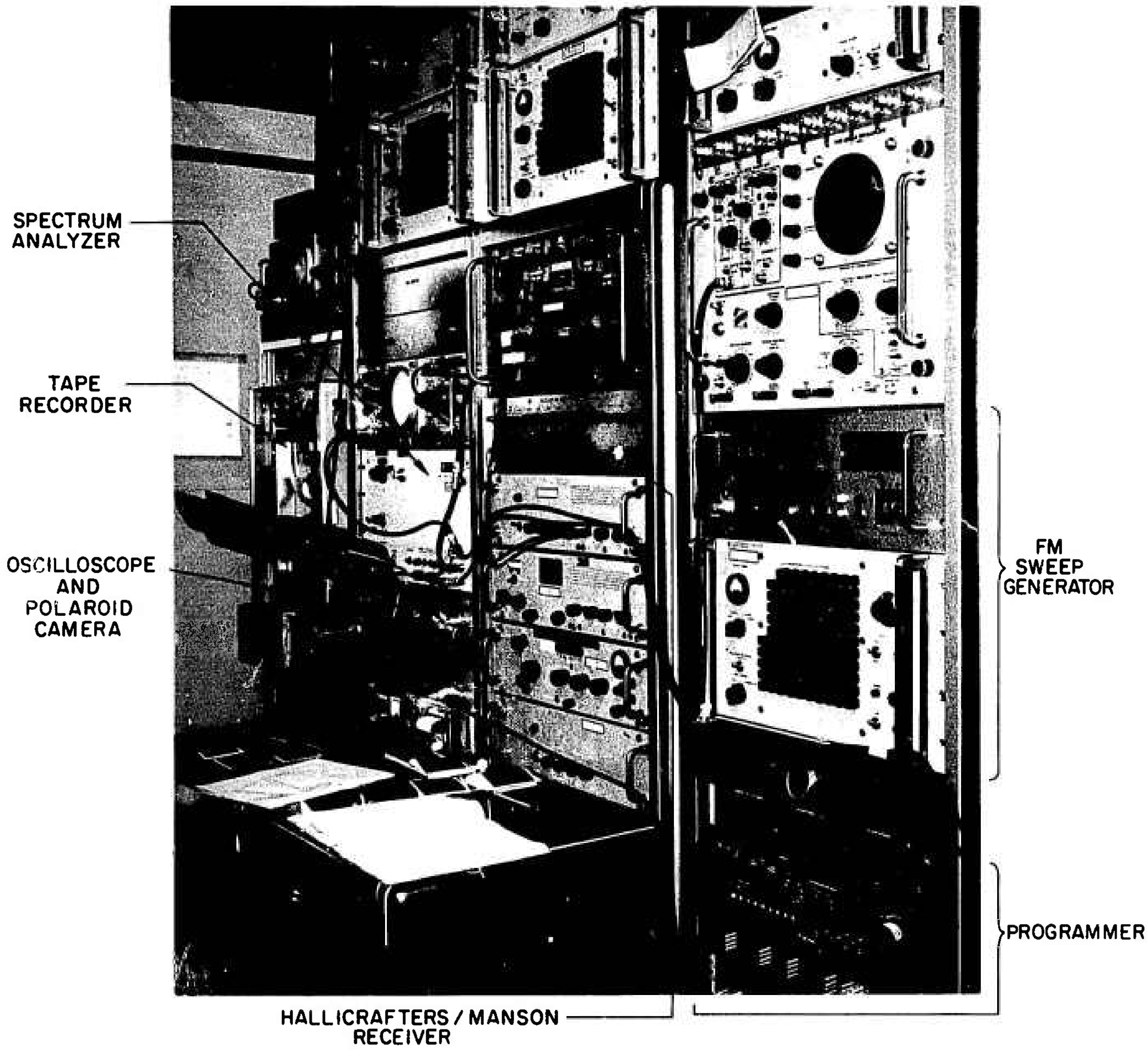


Fig. 19. CHIRP II SWEEP GENERATOR AND SOUNDER RECEIVING EQUIPMENT.

vertical input to produce an A-scan when desired; when the 35 mm film was pulled at high enough speed, the individual A-scans could be seen on the record.

B. Results

An example of an ionogram taken by the second method is shown in Fig. 21. The sweep rate was 1 MHz/sec; thus the 4- to 30-MHz ionogram was obtained in 26 sec. Despite the relatively low power (100 W), a considerable amount of detail is visible; there is very little evidence of interference from other spectrum users.

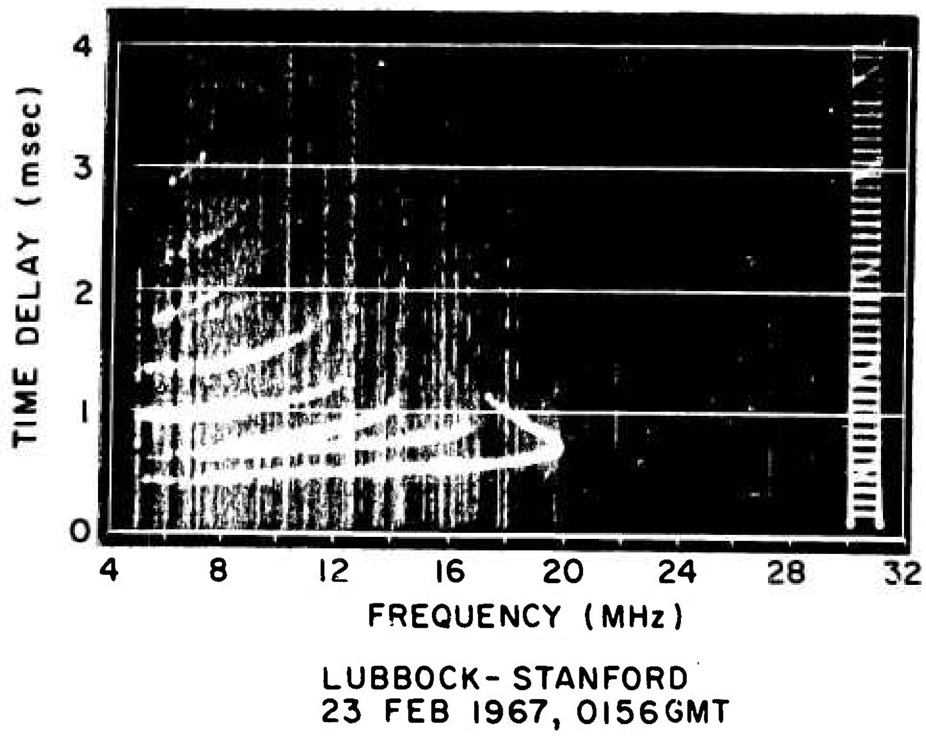


Fig. 20. EXAMPLE OF IONOGRAM USING MONITOR SCOPE AND POLAROID CAMERA.

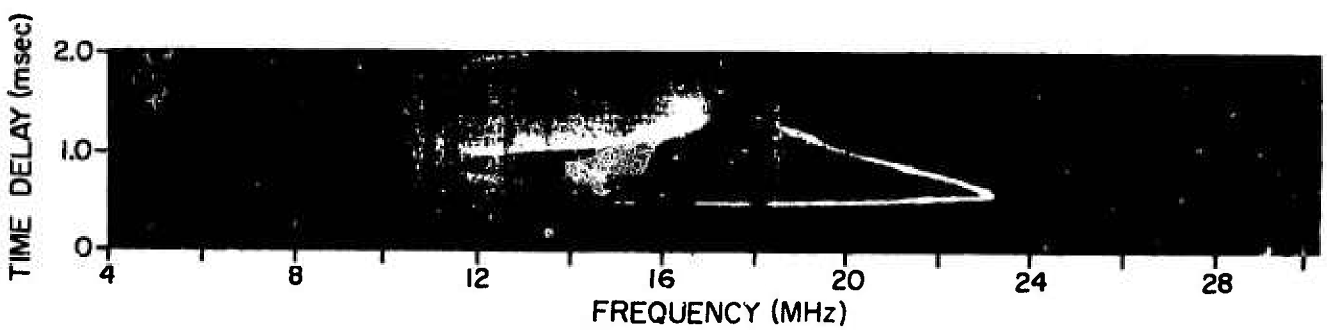


Fig. 21. IONOGRAM BY FM-CW METHOD. Transmitted power 100 W, sweep rate 1 MHz/sec, sweep time 26 sec.

Recording the received signal on tape gives additional flexibility in that the tape can be played back with varying amounts of speedup, thus varying the effective resolution of the system. (This technique is a means of circumventing the limitations of the spectrum analyzer used.) Figures 22 and 23 show ionograms produced by this technique. In each figure, the upper ionogram (a) was played back and analyzed with a 2X speedup. In (b) the tape speedup factor was 4X. Since the sampling rate of the spectrum analyzer was constant, only half as many traces

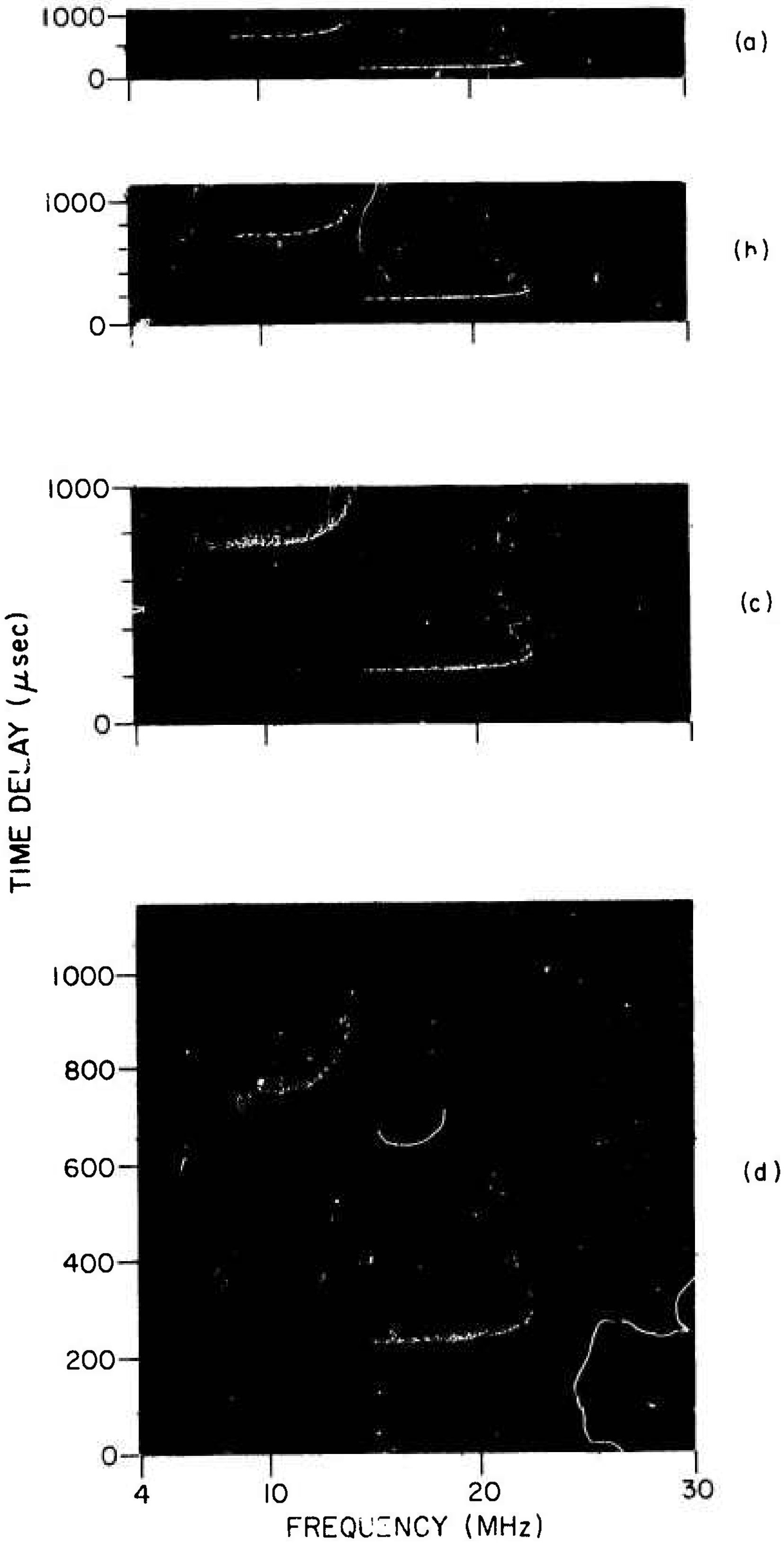


Fig. 22. IONOGRAM PROCESSED WITH TAPE SPEEDUP FACTORS OF (a), 2, (b) 4, (c) 8, and (d) 16.

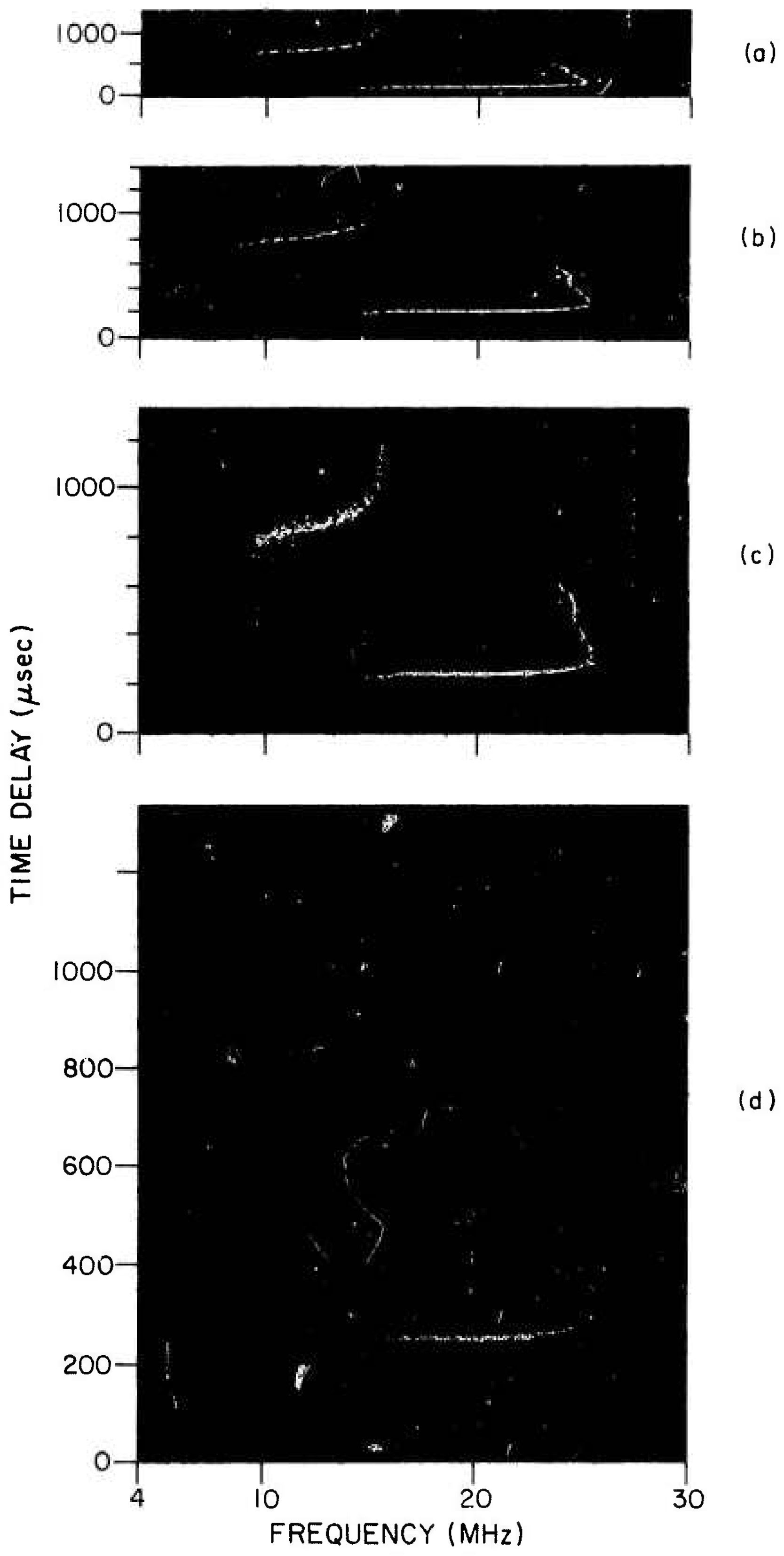


Fig. 23. IONOGRAM PROCESSED WITH TAPE SPEEDUP FACTORS OF (a) 2, (b) 4, (c) 8, and (d) 16.

were produced; thus the speed at which the film was moved was doubled to maintain the same size record on the horizontal (frequency) axis. By this means, horizontal (frequency) resolution was traded for increased vertical (time-delay) resolution. The individual scans for tape speed-up factors of 4X and greater are far enough apart that it is advantageous to feed the analyzer output signal also to the scope vertical channel; the resulting individual A-scans in (b) and (c) can be seen with a magnifying glass.

In Figs. 22c and 23c, the tape speedup was 8 times and the 35 mm film speed was correspondingly increased.

As the tape factor was increased beyond 2 times, with the 1-MHz/sec chirp rate used, the bandwidth of the played-back signal exceeded the 0- to 4-kHz range of the spectrum analyzer. It was therefore necessary to make several runs through each record, offsetting a conversion oscillator in the spectrum analyzer each time so that adjacent frequency ranges were analyzed, then to piece together the successive scans to make a continuous record of wider frequency range. This was done to make the ionograms of Figs. 22 and 23b, c, and d.

The records of Figs. 22d and 23d, made with 16 times speedup, show still greater resolution in time delay. The individual A-scans are clearly visible.

Figure 24 is a portion of an ionogram taken at 32 times speedup and optically enlarged, showing the F-2 lower ray. The finest detail in time delay shows a 3-dB resolution approaching 1 μsec.

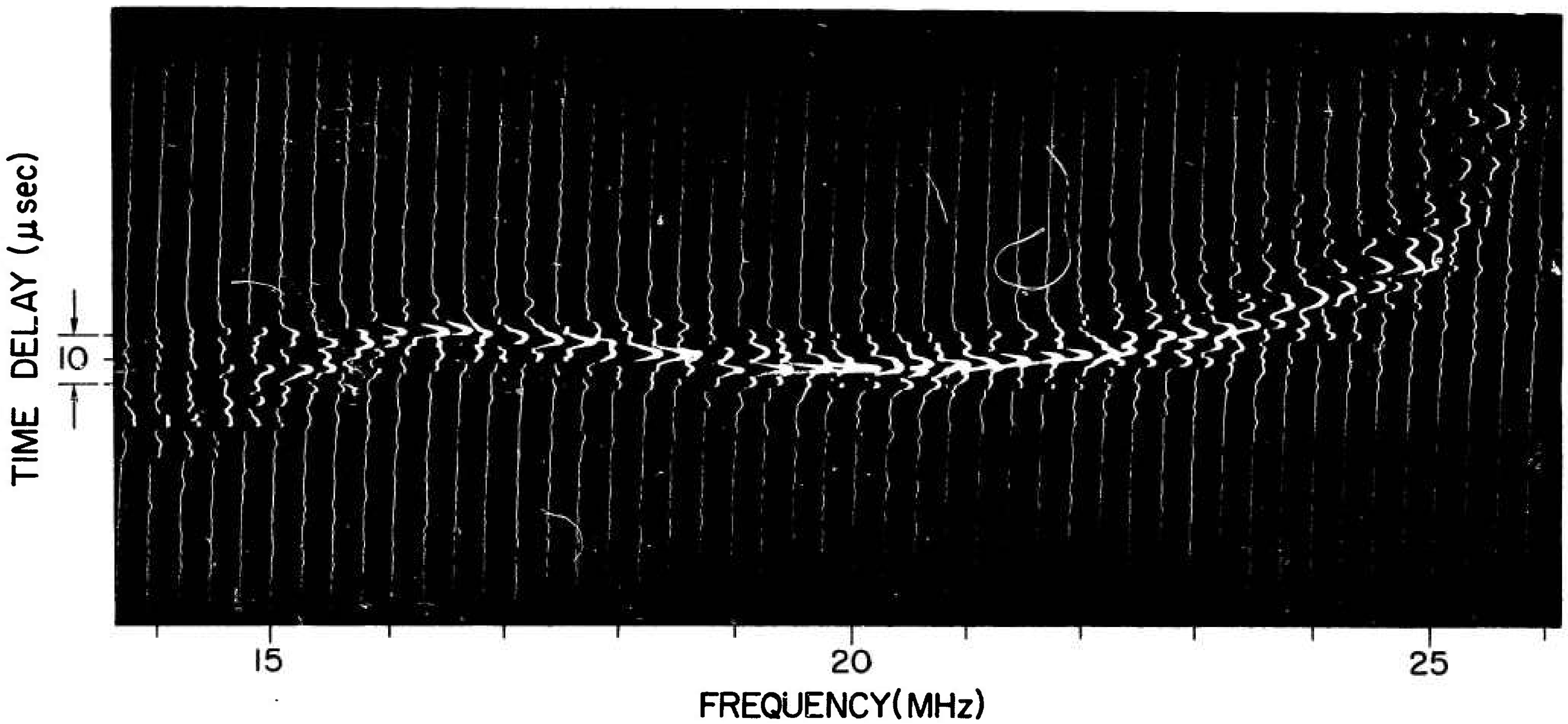


Fig. 24. IONOGRAM PROCESSED AT 32 TIMES TAPE SPEEDUP, SHOWING TIME DELAY RESOLUTION APPROACHING 1 μsec.

VI CONCLUSION

The objectives set up at the beginning of the Chirp II sweep-generator development program were essentially met. The completed equipment has the desired improvements in time-delay resolution, flexibility, ease of adjustment of sweep limits, sweep rates and starting frequencies, and a choice of sweep formats. Ability to switch quickly between fixed-frequency and sweep operations has been incorporated.

All performance criteria were met except sidelobe reduction, which was of the order of 20 dB (worst case) rather than the 25 to 35 dB expected on theoretical grounds. Some further improvement in this parameter is expected when the correct delay lines are installed.

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13. ABSTRACT

A frequency-modulation, continuous-wave HF ionospheric sounder ("Chirp I") was introduced in 1964. This equipment provided a considerable increase in system sensitivity over commonly used pulse sounders at a time-delay resolution of 10 μsec. Experience with this equipment revealed a need for still greater time-delay resolution capability, plus easily selectable sweep frequency limits and rates, and an ability to sweep with either positive or negative slopes.

In 1966, a second-generation equipment, "Chirp II," was developed. This report describes the differences between it and the "Chirp I" equipment. The salient characteristics of "Chirp II" are time-delay resolution of 1 μsec of less, great flexibility in selection of sweep limits and sweep rates, and a choice of several sweep formats. The sweep phase discontinuities have been reduced to a level more than adequate for most ionospheric sounding work, but further reduction is needed in some applications.

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