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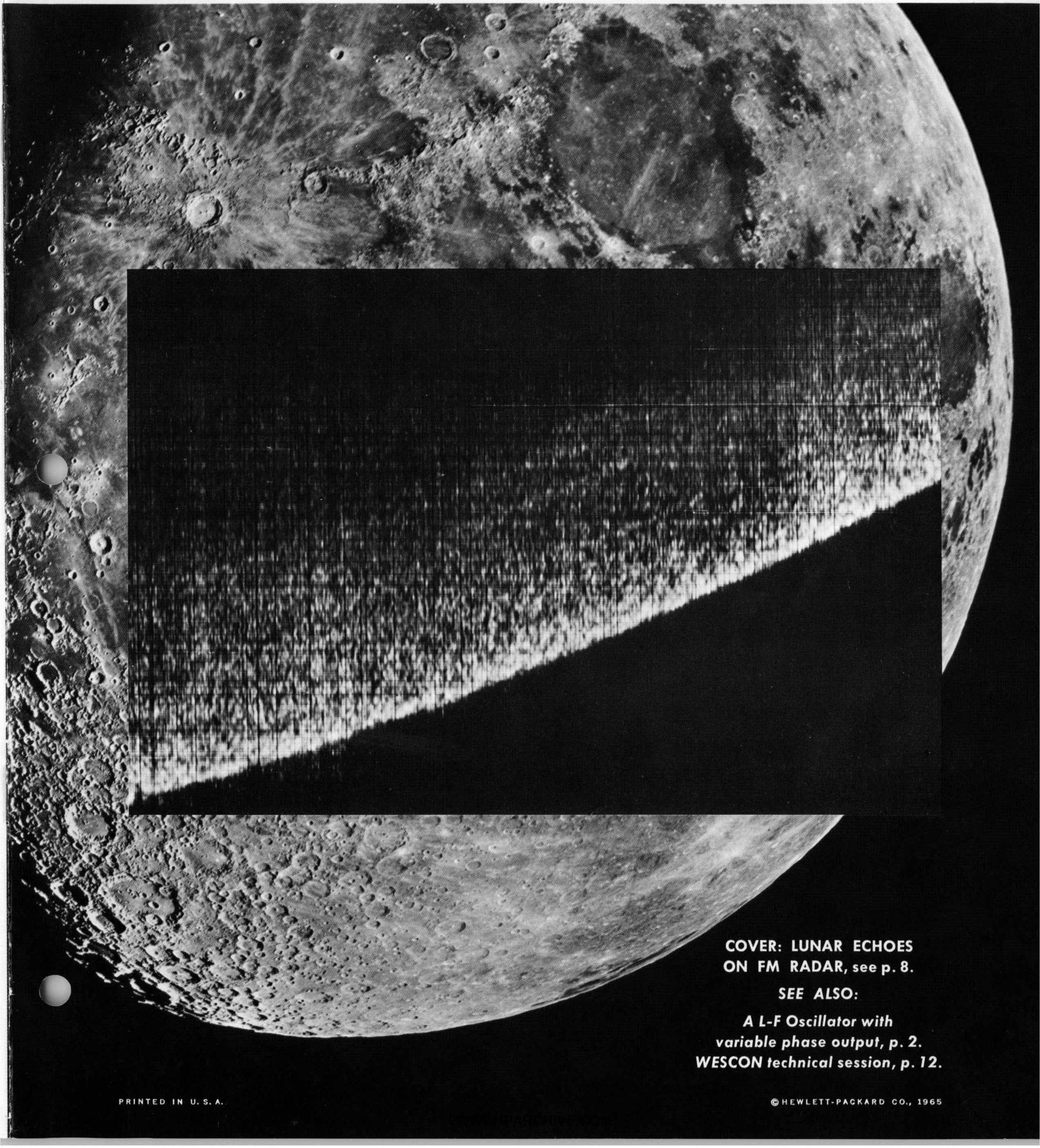

COVER: LUNAR ECHOES ON FM RADAR, see p. 8.

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A LOW-FREQUENCY OSCILLATOR WITH VARIABLE-PHASE OUTPUTS FOR GAIN-PHASE EVALUATIONS

A new l-f oscillator provides both sine and square outputs as well as adjustable-phase sine and square outputs over a range from 60 kc/s down to 0.005 c/s.

TEST SIGNALS at frequencies below the audio range are valuable in many applications, including the check-out of medical and geophysical equipment, servo-mechanism analysis, in process control testing, and as forcing functions for real time analogs of mechanical systems.

The usefulness of an oscillator for these low frequencies is enhanced if it provides a second output waveform that can be phase-shifted in known amounts with respect to the primary output. Among other applications, use of the phase-shifted output simplifies measurement of the phase shift encountered by a signal in propagating through a servo or process control system (see Fig. 3). The phase-shifted output is thus valuable for making gain-phase plots.

Accordingly, a new low-frequency oscillator has been developed, one that has a second output channel that can be phase-shifted continuously through a full 360° range. Although the new oscillator is intended primarily for low frequency work, it has a 12,000,000 to 1 frequency range extending from 60 kc/s down to 0.005 c/s or, with options, down to as low as 0.00005 c/s (5 hours for one cycle).

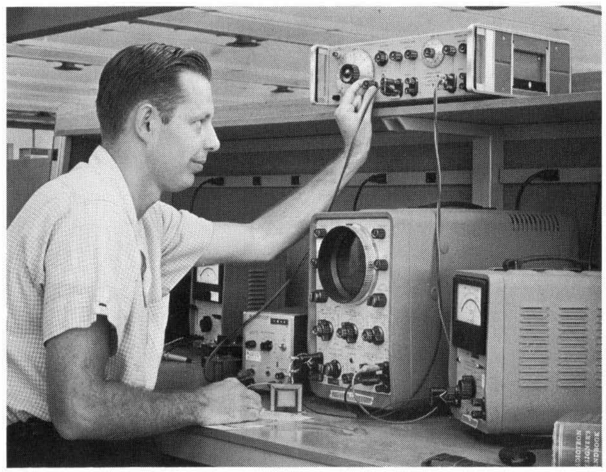


Fig. 1. –hp– Model 203A Variable Phase Function Generator simplifies measurement of gain-phase characteristics of circuits and devices, such as audio driver transformer shown here being checked throughout complete audio range with set-up diagrammed in Fig. 3 and explained in text. New generator has wide frequency range and variable phase outputs adjustable over full 360° range at any output frequency.

In addition to sine wave outputs, the new generator has separate square wave outputs on both the reference and phase-shifted channels. All four output signals, which have maximum amplitudes of 30 V p-p, are supplied simultaneously and all have individual 40-dB attenuators (see Figs. 2 and 5).

Output signals from the new oscillator, or Variable Phase Function Generator as it is formally known, are exceptionally clean. The rms total of all harmonic distortion, hum, and noise on the sine wave signals is more than 64 dB below the fundamental or, expressed in another way, is less than 0.06% of the fundamental (see Fig. 6). The new –hp– Model 203A Function Generator is thus well suited as a source of sine waves for critical tests of

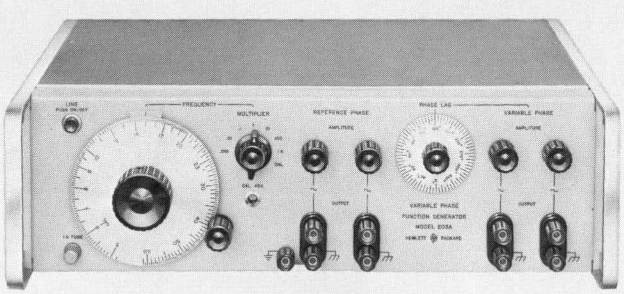


Fig. 2. *New –hp– Model 203A Variable Phase Function Generator has four output signals available simultaneously. Frequency range is from 60 kc/s at high end down to 0.005 c/s in seven overlapping ranges or, with options, down to 0.00005 c/s in nine ranges.*

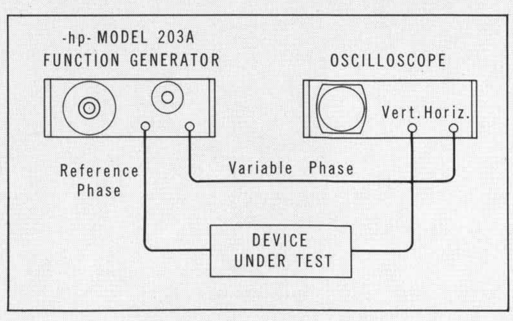


Fig. 3. *Set-up for phase-shift measurements using variable phase output achieves sharply defined null (see Fig. 9).*

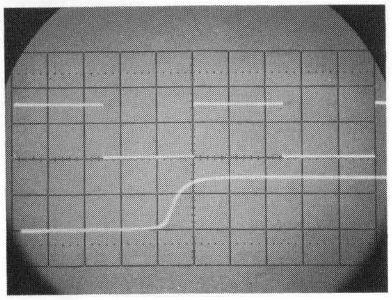
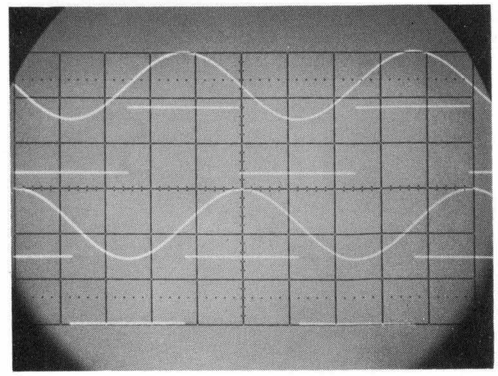


Fig. 4. *Reference-phase square-wave output of Function Generator as viewed on oscilloscope triggered by variable-phase square-wave output. Use of variable-phase output as trigger allows reference phase signal to be positioned on CRT as desired. (Upper trace: 20 μs/cm; lower trace: 0.2 μs/cm.)*

Fig. 5. *Variable Phase Function Generator supplies four waveforms simultaneously, as shown by this oscillogram made by four-channel oscilloscope sensing all four outputs at same time. One sine wave and one square wave output (lower two traces) may be phase-shifted together with respect to reference sine- and square-waves (upper two traces). Each output has separate 40-dB attenuator and is capable of 30V peak-to-peak maximum voltage.*



audio as well as sub-audio equipment. The excellent waveform purity is maintained throughout the entire frequency range of the instrument.

The square wave tops are flat and rise and fall times are less than 0.2 μs with less than 5% overshoot (Figs. 4 and 5). Square-waves of this quality are well suited for transient response testing of high-grade audio and other low frequency equipment.

Frequency accuracy of the new oscillator is better than ±1% at all frequencies and the instrument may be quickly calibrated to local power lines operating within a 60 to 1000 c/s frequency range. The instrument is transient-free and permits rapid changes of frequency without requiring a "settling down" period.

APPLICATIONS OF A VARIABLE PHASE OUTPUT

The phase-shifted output has a number of useful applications. Obviously, the phase-shifted square-wave is useful as a sweep trigger for an oscilloscope, while either the sine or square wave output of the reference channel serves as a test signal for the device under examination (Fig. 4). Any portion of the test signal waveform may then be positioned on the oscilloscope screen with the PHASE LAG dial of the new generator.

The PHASE-LAG dial may also serve as a "manual scan" control for the Moseley Model 101 Waveform Translator, a useful arrangement for making X-Y plots of tests conducted at audio frequencies. The Waveform Translator is a sampling device designed to translate repetitive high-frequency oscilloscope T-Y patterns to a lower frequency range for graphic recordings. As shown in the typical application diagrammed in Fig. 7, the Waveform Translator may be triggered by the variable-phase square-wave output of the Function Generator. Then by turning the PHASE-LAG dial manually, both the X and Y waveforms may be scanned by the sampling pulses at a rate compatible with X-Y recorder operation.

Use of the variable phase signal itself as the test signal provides an operating convenience at very low frequencies, where one cycle may last for several minutes. Any portion of the waveform — for instance, the step transition in the square wave serving as a stimulus for a real-time analog computer system — may be quickly selected simply by turning the PHASE-LAG dial to reposition the waveform.

Push-pull outputs, either sine wave or square wave, are easily obtained by setting the PHASE-LAG dial to 180° and using both outputs with the "common" connectors as the center tap of a balanced system. The common connectors are floated from chassis ground, permitting offsets up to 500 V dc.

Complex rectangular waveforms are obtained simply by summing the reference phase and variable phase square-wave outputs. Individual output attenuators and the controllable phase shift enable the generation of a wide variety of waveforms (Fig. 8).

A technique using the variable phase output for phase-shift measurements

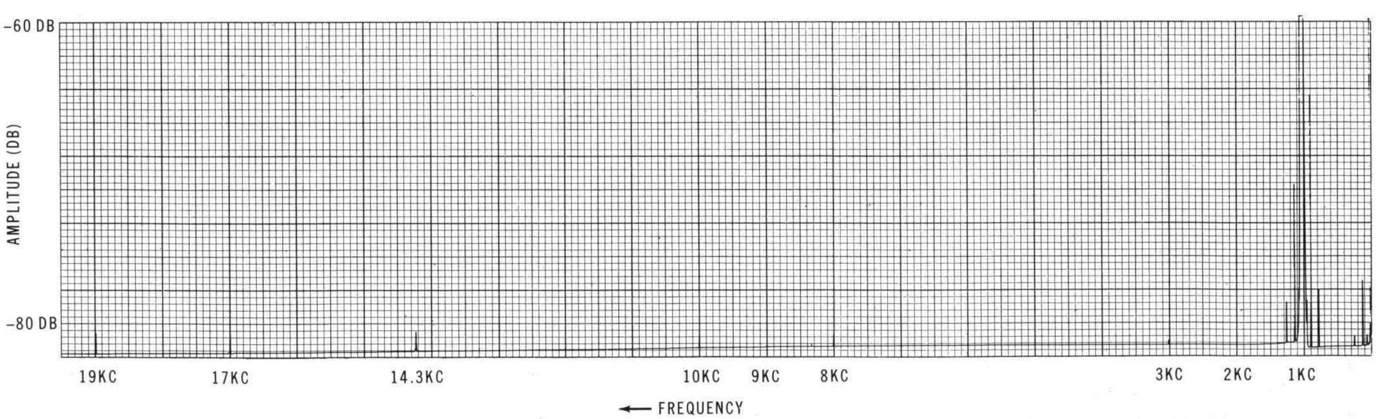


Fig. 6. *Typical harmonic content in sine-wave output of Model 203A Function Generator at 1 kc/s shows low distortion in output wave.*

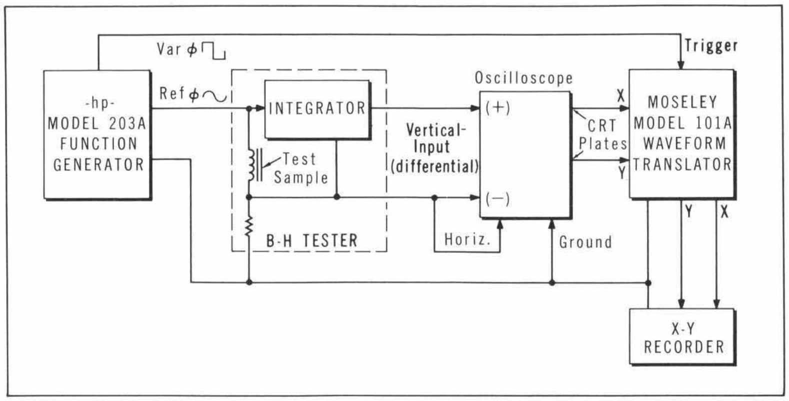


Fig. 7. *Instrument set-up for making X-Y plots of B-H curves obtained at audio frequencies.* Phase Lag *dial of Variable Phase Function Generator permits manual control of both X and Y waveform scanning by Waveform Translator. Waveform Translator repetitively samples both X and Y signals in oscilloscope, translating waveforms to equivalent low frequency signals for operating X-Y recorder. (Some measurements may require addition of power amplifier at "Ref φ" output.)*

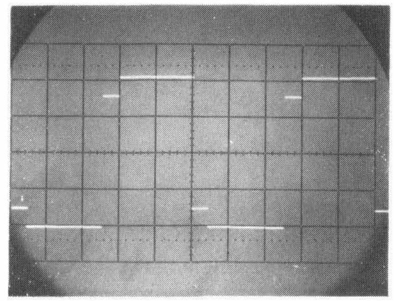


Fig. 8. *Complex rectangular waveforms obtained by summing Function Generator square-wave outputs in resistive network.*

is outlined in the block diagram of Fig. 3. The reference phase supplies the sine-wave test signal for the device under test while the output of the device is applied to the oscilloscope vertical channel. The variable phase feeds the horizontal input of the oscilloscope directly. In making the measurement, the operator adjusts the PHASE-LAG dial to close up the displayed ellipse to a straight line with positive slope, which achieves a sharply defined null (Fig. 9). The phase shift in the system under test then is read directly from the setting of the PHASE-LAG dial. No calculations involving the parameters of the displayed ellipse or the settings of the amplitude controls are required with

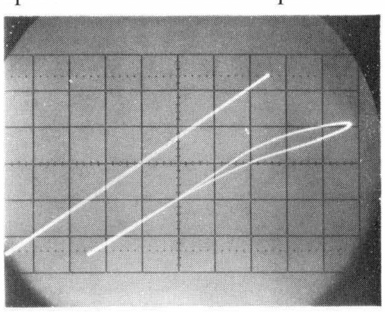


Fig. 9. *Double exposure oscillogram illustrates use of variable-phase output in phase measurements, as performed by set-up diagrammed in Fig. 3. Oscillator* Phase Lag *dial has been adjusted to close up Lissajous ellipse in left-hand pattern. Null thus achieved is sharply defined. In right-hand pattern, vertical signal is distorted by flattening of positive peaks. Accurate phase-shift measurement can be realized nevertheless by adjustment of* Phase Lag *dial to close up lower part of ellipse.*

this measurement. The method has a further advantage in that the measurement is relatively unaffected by distortion that may be introduced in the test signal by the device under test, as shown by the second pattern in Fig. 9.

Phase-shift measurements may also be made by using the leading and trailing edges of the variable-phase square wave as oscilloscope intensity modulation markers for establishing phase relationships, as shown in Fig. 11.

Amplifier distortion can be studied readily with the help of a variable phase output. The reference phase serves as the test signal in the usual manner and the variable phase is used to null out the test signal at the amplifier output, either in a resistive summing network, or at the minus input of an oscilloscope with differential input. Through proper adjustment of the phase and amplitude of the variable phase signal, the fundamental of the test signal can be nulled out, leaving only the distortion products for display. This technique is particularly valuable when the distortion components are very small.

THEORY OF OPERATION

The requirement for variable phase output led to the use of beat frequency techniques in the new Function Generator. This allows the phase shifting device to be placed in the fixed frequency channel where it always operates at the same frequency. Calibration accuracy of the phase-shifter with respect to the reference channel is therefore maintained at all times regardless of the selected output frequency.

Where very low frequencies are concerned, however, beat frequency oscillators have been prone to excessive drift and to other instabilities and distortions caused by the tendency of two oscillators to "lock in" when tuned to very nearly the same frequency. These problems were overcome in the new low-frequency oscillator by the use of techniques that have been identified with frequency synthesis.

Operation of the instrument on the highest frequency range (5 to 60 kc/s), is shown by the simplified diagram of the reference channel in Fig. 10. The

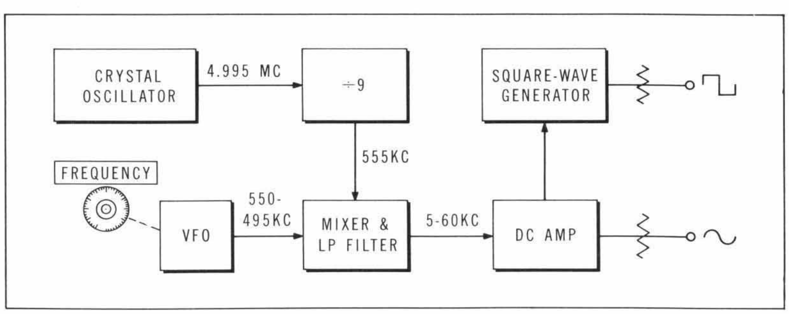


Fig. 10. *Simplified block diagram of reference phase channel in -hp- Model 203A Function Generator while operating on highest frequency range (5–60 kc/s).*

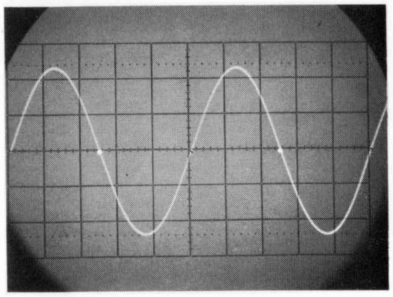


Fig. 11. *Phase-shift measurement by Z-axis modulation of test waveform by differentiated variable phase square-wave output. Differentiated square wave brightens CRT trace on negative-going steps and blanks trace on positive-going steps. Variable phase square wave is in phase (or 180° out of phase) with waveform applied to vertical deflection channel when blanking and brightening spots are aligned along horizontal line on CRT trace.* Phase Lag *dial on Function Generator then reads phase shift.*

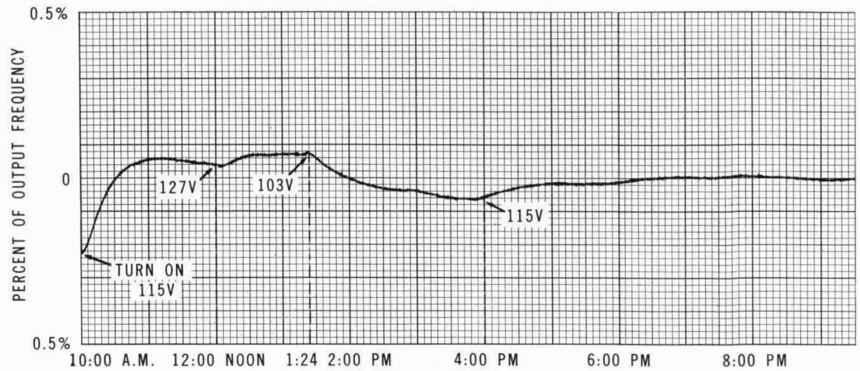


Fig. 12. *Graph of frequency stability of –hp– Model 203A Function Generator shows effects of warm-up and high and low line voltages. Instrument was first turned on at 10 am at start of graph. At 12 o'clock, line voltage was raised 10% above normal and then reduced to 10% below normal at 1:24 pm. Line voltage was returned to normal at 4 pm and from then on was subject to normal line voltage fluctuations. Peak-to-peak short-term stability (determined by width of trace) is better than 1 part in 10^4.*

fixed frequency is derived by dividing down a crystal oscillator frequency by a factor of 9, the factor of 9 being required by the frequency changing operations used for the lower frequency ranges.

The fixed-frequency output (FFO) from the $\div 9$ divider is heterodyned with the variable frequency oscillator (VFO) signal in a doubly-balanced mixer to derive the output signal. Because the sum frequency and other products incidental to the process are so far removed from the difference frequency, they are easily removed by filtering.

The square-wave output is derived by applying the output sine wave to a clipping amplifier and improving the risetime of the resulting square-wave in a dc-coupled regenerative circuit.

Frequency stability (see Fig. 12) in the new Function Generator has been achieved by careful design of the VFO. The oscillator transistor is operated at low collector current for good short-term stability while long-term and temperature stabilities are achieved by stringent specification of the frequency-determining components (L and C).

DECADE FREQUENCY DIVISION

The lower frequency ranges are derived by frequency arithmetic, as outlined in the block diagram of Fig. 13. In describing the technique, we may consider the fixed frequency (555 kc/s) as f_o, the crystal oscillator frequency as $9f_o$ and the VFO frequency as $(f_o - \Delta f)$. Obviously, the beat frequency between the fixed and VFO frequencies is Δf.

The VFO output is applied to the balanced mixer in a *Decade Module* where it beats with the crystal oscillator frequency ($9f_o$). Tuned circuits select the sum frequency component $[9f_o + (f_o - \Delta f)]$ and this frequency is divided by 10 to obtain an output $(f_o - \frac{\Delta f}{10})$ which, of course, differs from the fixed frequency f_o by $\Delta f/10$. This frequency serves as the VFO signal for the instrument when the frequency multiplier switch is set to $\times 100$, the next highest range, as shown in the diagram. The output frequencies covered in this range (0.5–6 kc/s) thus become exactly one-tenth of those covered in the highest frequency range (5–60 kc/s).

Although the frequency range has been divided by 10, the relative stability of the generator output on this range is the same as on the highest range. This is because any absolute instability in the VFO is also divided by 10 in the decade module.

The output of the first decade module is also supplied to the mixer of the second decade module, as shown in the diagram of Fig. 14, where it is added to the crystal oscillator frequency ($9f_o$). The sum frequency output of the mixer in this decade module is $(10f_o - \frac{\Delta f}{10})$, and it is divided by 10 to obtain $(f_o - \frac{\Delta f}{100})$. This frequency is

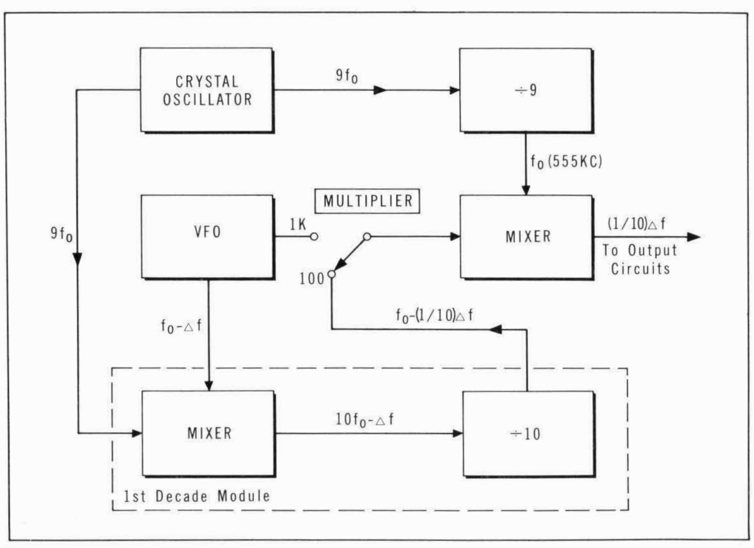


Fig. 13. *Block diagram of frequency generating circuits while operating on next to highest frequency range (0.5–6 kc/s).*

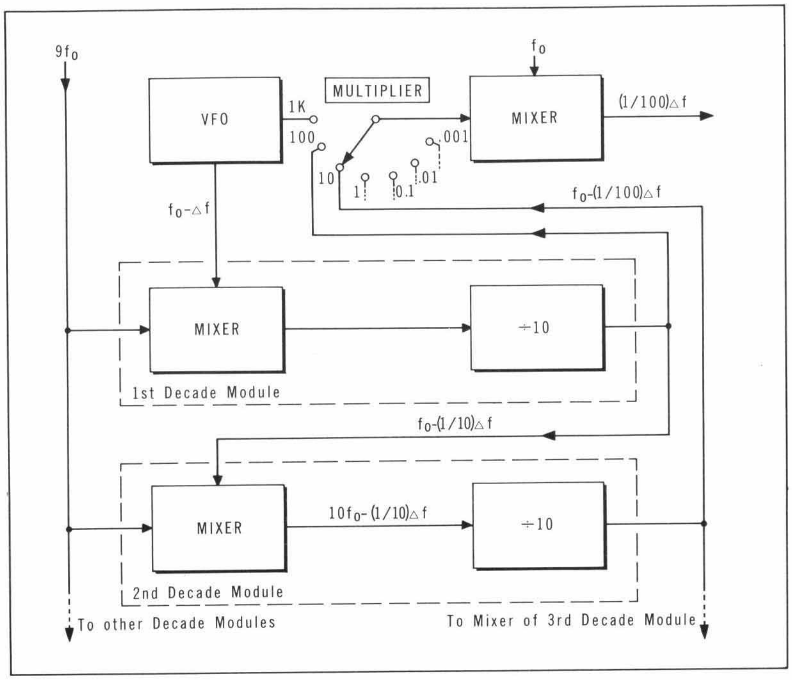


Fig. 14. *Block diagram of frequency-generating circuits on lower frequency ranges.*

used as the VFO signal for the third highest range (50–600 c/s), once again reducing the tunable output frequency range by a factor of exactly 10. Output stability likewise is preserved.

This add-and-divide-by-10 process is repeated four more times in identical decade modules to obtain the lower frequency ranges. Space was left in the instrument for two additional modules to extend the frequency range by two optional lower decade ranges.

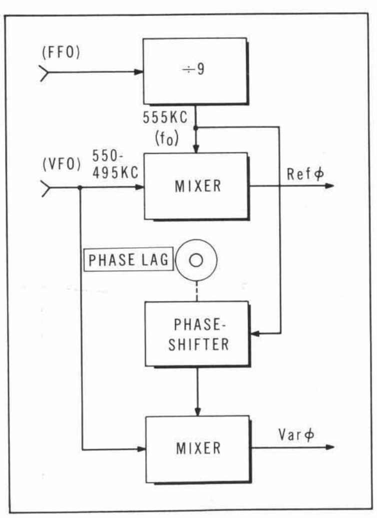


Fig. 15. *Variable phase channel has phase shifter in fixed frequency path but otherwise is identical to reference phase channel.*

Note that none of the frequency products within the instrument are very close to either the crystal oscillator or variable frequency oscillator outputs. These oscillators therefore are not prone to instabilities or distortions that would result from a tendency toward "locking in," as usually happens when two oscillators operate at very near the same frequency.

PHASE SHIFT CIRCUITS

An accurate goniometer, placed in the 555 kc/s fixed frequency (FFO) signal path leading to the variable phase output, develops the output that is phase-shifted with respect to the reference output (Fig. 15). The goniometer has two stator field windings, oriented at right angles and excited at 90° with respect to each other, and a pick-up coil (armature) that can rotate within the stator fields. The phase of the signal induced in the pick-up coil depends on the coil's angular position. Careful design and adjustment of the stator fields results in a linear relationship between electrical phase and angular position (Fig. 16).

The goniometer pick-up coil is attached to a shaft turned by the front panel PHASE-LAG dial. Slip rings take the signal from the pick-up coil and the shaft may thus be continuous-turning, allowing any amount of phase shift to be cranked into the variable phase signal.

The amount of phase shift introduced into the fixed frequency path of the variable phase channel is translated into an equivalent phase shift in the heterodyned output frequency.

Following the goniometer and associated parts, both the sine and square wave circuits of the Variable Phase channel are identical to those in the Reference Phase channel. A minimum differential phase shift therefore exists between the two channels, thus preserving the phase relationship established by the goniometer.

DC OUTPUT CIRCUITS

The mixers which produce the difference frequency between the FFO and VFO are dc-coupled, doubly-balanced low-distortion modulators. Any dc drift caused by temperature changes or voltage drifts in the circuits could have a rate of change comparable to the very low frequency outputs, and would thus be a source of distortion. To reduce this, each output amplifier has a differential input. A "common-mode" correction signal, corresponding to drift sensed in the modulator, is applied to the out-of-phase input of

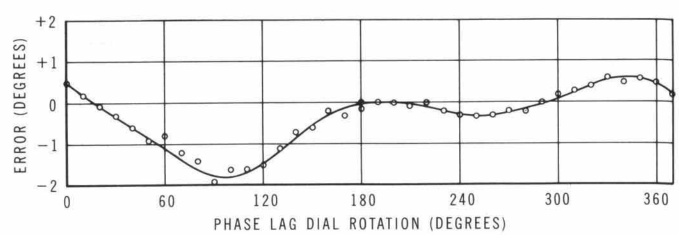


Fig. 16. *Plot shows small departure from linearity of* Phase Lag *dial of typical Model 203A Variable Phase Function Generator. Errors were determined at output of about 300 c/s by comparing* Phase Lag *dial reading to phase difference as measured between Reference and Variable channels with time interval meter.*

the differential amplifier to obtain dc stability of a high order.

AMPLITUDE STABILITY

Automatic level control in the RF amplifiers (not shown) which precede the output mixer hold the mixer inputs at a constant amplitude irrespective of the tuning of the instrument. The mixer itself operates with constant gain and the output amplifier has 60 dB of feedback for gain stability, as well as low distortion. The output frequency response therefore is level within $\pm 1\%$, referenced to the output level at 1 kc/s, as shown in Fig. 17. The long term output amplitude stability, including warm-up drift and power line variations of $\pm 10\%$, likewise is better than $\pm 1\%$ (± 0.1 dB) and the instrument retains this stability despite changes in time, temperature, or frequency. In a laboratory environment, amplitude stability is typically within 1 part in 10^4.

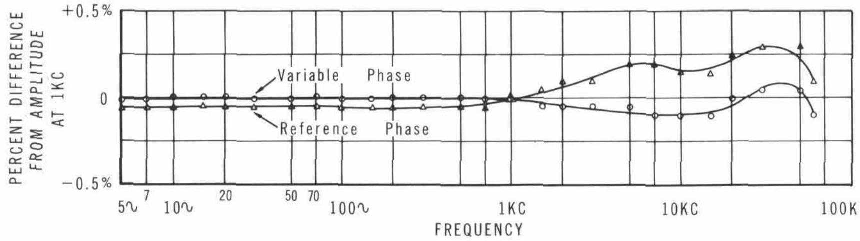


Fig. 17. *Amplitude vs frequency for typical –hp– Model 203A Variable Phase Function Generator.*

CALIBRATION

The instrument frequency accuracy is verified quickly by setting the MULTIPLIER switch to the CAL position. This feeds both the power line ac and one of the square wave outputs to the pilot lamp, which then flickers at twice the resulting beat rate. With the dial set at "6" for 60 c/s power lines, or at other positions for other line frequencies or subharmonics of line frequencies, the VFO frequency may be trimmed with a front panel screwdriver control to reduce the beat frequency to zero, thereby calibrating the instrument to the power line frequency. This one adjustment calibrates the instrument for all ranges, because of the exact $\times 10$ relationship between ranges.

ACKNOWLEDGMENTS

Most of the basic design ideas in the new –hp– Model 203A Low Frequency Function Generator were contributed by Donald Norgaard. The author also wishes to acknowledge the contributions of Michael Farrell, William Voisinet, and Paul Stoft.

—Richard Crawford

SPECIFICATIONS
–hp–
MODEL 203A
VARIABLE PHASE FUNCTION GENERATOR

FREQUENCY RANGE: 0.005 c/s to 60 kc/s in seven decade ranges.*

DIAL ACCURACY: $\pm 1\%$ of reading.

FREQUENCY STABILITY: Within $\pm 1\%$ including warmup drift and line voltage variations of $\pm 10\%$.

OUTPUT WAVEFORMS: All waveforms available simultaneously. All outputs have common chassis terminal.

REFERENCE PHASE: Sine Wave and Square Wave.

VARIABLE PHASE: Sine Wave and Square Wave, continuously adjustable in phase from 0–360°.

PHASE DIAL ACCURACY: $\pm 5°$ sine wave; $\pm 10°$ square wave.

OUTPUT SYSTEM: Direct coupled output is isolated from ground and may be operated floating up to ± 500 V dc.

MAXIMUM OUTPUT VOLTAGE: 30 volts peak-to-peak open circuit for sinusoidal and square waveforms.

OUTPUT POWER: 5 volts rms into 600 ohms (40 mW); at least 40 dB continuously adjustable attenuation on all outputs.

DISTORTION: Total harmonic distortion, hum and noise >64 dB below fundamental (<.06%).

AMPLITUDE STABILITY (with respect to frequency): $\pm 1\%$ referenced to 1 kc/s.

SQUARE WAVE RESPONSE:
RISE AND FALL TIME: <0.2 μs.
OVERSHOOT: <5%.

POWER: 115 or 230 V $\pm 10\%$, 50 to 1000 c/s, approximately 25 watts.

SIZE: 5¼ in. high x 16¾ in. wide x 11½ in. deep.

WEIGHT: Net, 19 lbs. 4 oz. (8,66 kg). Shipping, approximately 25 lbs. (11,25 kg).

PRICE:
MODEL 203A: $1,200.00
OPTION 01: Add $40.00
OPTION 02: Add $80.00

Prices f.o.b. factory.
Data subject to change without notice.

* Two lower ranges of 0.0005 c/s (option: 01) and 0.00005 c/s (option: 02) are available on special order.

DESIGN LEADERS


Richard Crawford Donald E. Norgaard

Dick Crawford joined Hewlett-Packard as a developmental engineer in 1960. In addition to engineering the 203A Variable Phase Function Generator, he has worked on the 310A Wave Analyzer, the 3460A Digital Voltmeter, and on photoelectric switching devices. At present he is with the –hp– Advanced Research and Development Laboratories where he is concerned with circuitry for display systems.

Following 2 years service in the U. S. Army as a microwave technician, Dick attended the California State Polytechnical College where he earned a BSEE just prior to joining –hp–.

Following graduation from Rice Institute with a BSEE degree, Don Norgaard became engaged in television studio equipment design and as part of this activity, he participated in the field trials leading toward establishment of the NTSC TV standards. He also served on other industry committees concerned with TV equipment and transmission standards.

During World War II, Don designed naval gunfire control radar. Following the war, he engaged in research on communications systems, with particular emphasis on single-sideband modulation where he made significant contributions* that have led to the realization of practical SSB systems.

Don joined Hewlett-Packard as a development engineer in 1957 and has been concerned with the design of numerous instruments, primarily in the audio-video field, with major participation in the 425A Micro-Voltammeter and 738A Voltmeter Calibrator designs among others. He also contributed circuit ideas for the Dymec 2800A Quartz Thermometer and was Project Leader on the –hp– Model 412A DC Voltmeter. At present, he is concerned with electronic designs in systems for chemical analysis at the -hp- Mechrolab Division.

Don is a member of Phi Beta Kappa and Tau Beta Pi. He is an active amateur radio operator and was chief designer of the OSCAR III Translator Satellite which was orbited by the Air Force on March 9, 1965.

* Donald E. Norgaard, "The Phase-Shift Method of Single-Sideband Generation," and "The Phase-Shift Method of Single-Sideband Reception," **Proc. IRE**, Vol. 44, No. 12, Dec. 1956.

EXTRATERRESTRIAL AND IONOSPHERIC SOUNDING WITH SYNTHESIZED FREQUENCY SWEEPS

by

G. H. Barry and R. B. Fenwick

Radioscience Laboratory, Stanford University

INTRODUCTION

Properties of the earth's upper atmosphere and of interplanetary space are currently the subject of intensive investigation. Before the advent of rockets and space vehicles, radio signals had been used for many years to probe this region and they still provide a convenient and inexpensive exploring means even though direct measurements now are available. One organization engaged in research using radio and radar techniques is the Radioscience Laboratory of Stanford University whose scientists succeeded this year in applying the well-known advantages of "Chirp" radar to the problems of ionospheric and extraterrestrial radio sounding. The advance was made possible through a novel application which used the -hp- Frequency Synthesizer for synthesizing highly-accurate frequency-sweep signals. This application is described in the accompanying article by Drs. George Barry and Robert Fenwick of the Radioscience Laboratory at Stanford.

Most of our present understanding of the ionized upper atmosphere has been obtained from pulse radar studies. One of the most useful measurements of the ionosphere is a profile of electron density versus height above a specified point on the earth. Vertical-incidence radio sounders (ionosondes) have for many years been used to give "virtual height"-versus-frequency records from which the electron density may be deduced. The operating frequency of the ionosonde is varied slowly across the hf (3–30 MHz) frequency range, and the sounder records the apparent height at which the ionosphere reflects signals. Reflection occurs where the electron plasma frequency equals the probing radio frequency, so the measurement gives, for each frequency, the height at which the corresponding electron density exists. Group-time-delay versus frequency "ionograms" are obtained by photographing intensity-modulated displays of the received pulse signals. An example is shown in the lower record of Fig. 2.

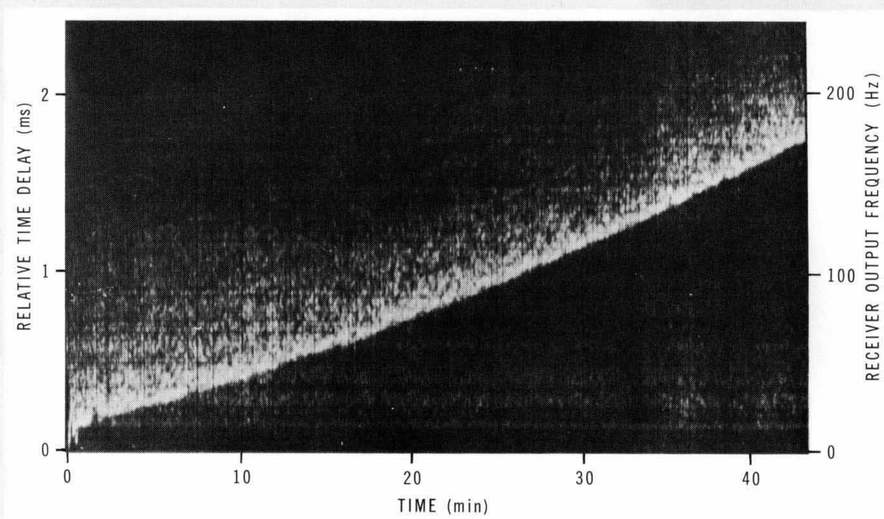


Fig. 1. *Lunar echoes obtained at Stanford University by FM radar operating with carrier sweep of 24.9 to 25.0 MHz. Vertical scale represents normalized time delay between transmitted and received signals as derived from frequency difference between signals. Curve of bright edge represents change of distance between earth and moon during measurements. Faint lines running parallel to bright edge result from delayed echoes scattered by prominences on moon.*

The ionosonde measures the ionosphere directly overhead, and from this information it is possible to predict the behavior of radio signals propagating between separated points on the earth's surface. A more direct measurement of propagation conditions over actual communications paths is provided by oblique sounding. With this technique, the equipment performs the same functions as the vertical-incidence sounder — transmitting and receiving pulse signals over the entire hf frequency range — but is more complex since the transmitter and receiver are separated geographically but must operate in synchronism. From oblique-sounder records, the hf radio communicator can determine which frequencies are propagating, and how well.

Certain characteristics of the ionosphere, such as ionospheric irregularities and layer tilts, are best studied by a technique which gives an overall view of a large area, rather than localized information such as is provided by vertical-incidence or oblique sounders. While networks of sounders have been used, an alternative technique is the use of ground backscatter sounding to survey large areas from a single location. A backscatter sounder makes

AUTHORS

Dr. George H. Barry (left), Senior Research Associate, and Dr. Robert B. Fenwick, Research Associate, of the Stanford University Electronics Laboratories, Radioscience Lab.

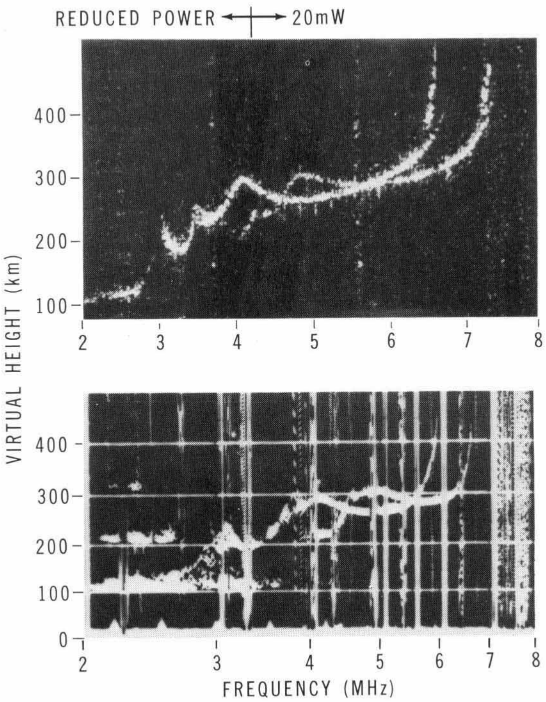


Fig. 2. *Ionograms obtained during same epoch by "Chirp" radar operating CW with output power of 20 mW (upper photo) and conventional 2-kW, 50-μs pulse sounder (lower photo). Broad vertical stripes in pulse sounder ionogram, only faintly visible in FM radar, result from interference.*

use of the fact that radio energy, after ionospheric reflection, is scattered when it strikes the earth. Some of the ground-scattered energy can thus be detected at a receiver located at or near the transmitter. The time delay defines the range to the scattering region, and changes in ionospheric refraction and focusing are apparent on records of the backscatter echoes.

Radar measurements are not restricted, of course, to the region beneath the earth's ionosphere. Echoes are obtainable from the sun, moon, and planets as well. The electron content of space beyond the ionosphere can be inferred from the behavior of radio echoes obtained from the moon. For this purpose, and also to explore the scattering behavior of the lunar surface, the Center for Radar Astronomy at Stanford has conducted radar measurements of the moon at 25 and 50 MHz (Fig. 1).

CHIRP SOUNDING

The ionospheric and lunar measurements described in the preceding have been obtained, or at least attempted, in the past using pulse sounders. The possibility of improving record quality through the application of frequency-sweep or 'Chirp'-radar techniques had been recognized for several years. The Chirp radar uses a linear-frequency-modulated pulse which results in two significant advantages—higher average power and less vulnerability to narrow-band interference. The Chirp technique had not, however, been applied in hf radio sounding because of the lack of a sufficiently accurate frequency-sweep signal. A newly-developed waveform synthesis technique now overcomes this problem and produces records of much improved quality, as shown in Figs. 2 and 8.*

Fig. 3 shows a frequency/time representation of pulse and Chirp signals, together with sounding echoes from these signals. For the same bandwidth (as shown), the two systems have identical time resolution. The FM sweep, or Chirp, achieves much higher average power because the duration of the frequency-sweep signal can be made arbitrarily long.

Chirp radar's increased immunity to narrow-band interference is important in this application because of the apparent high percentage of occupancy of the hf radio spectrum (Fig. 4). In sounding work a time resolution to an accuracy of 10 to 50 μs is generally desirable, implying pulse receiver bandwidths of 20 to 100 kHz. With an FM sweep system, though, the receiver bandwidth may be made arbitrarily small simply by slowing down the sweep rate (provided, of course, that an arbitrarily-accurate frequency sweep is

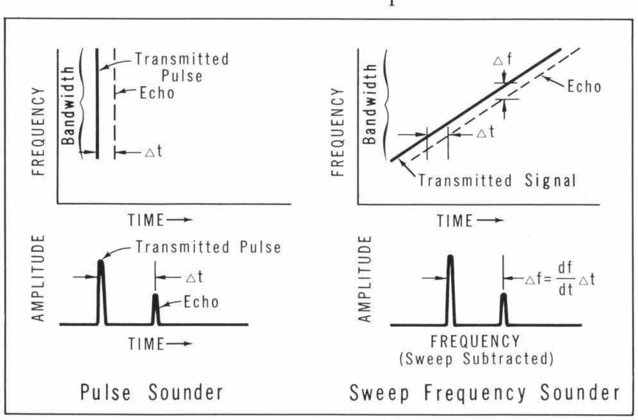


Fig. 3. *Frequency-time representation of pulse and Chirp radar signals. Time delay Δt of Chirp radar echo is derived from frequency difference Δf between transmitted and received signals.*

* The technique was developed at Stanford under support from the Advanced Research Projects Agency through the Office of Naval Research, Contract Nonr 225(64).

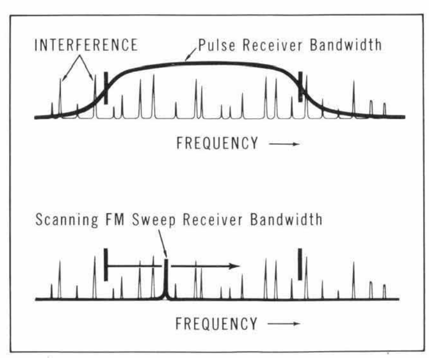


Fig. 4. *Receiver bandwidths are shown here superimposed on typical interference spectra encountered in hf communications band. Pulse receiver requires broad bandwidth for range resolution. Bandwidth of scanning FM (Chirp) receiver can be made arbitrarily narrow with no loss in resolution.*

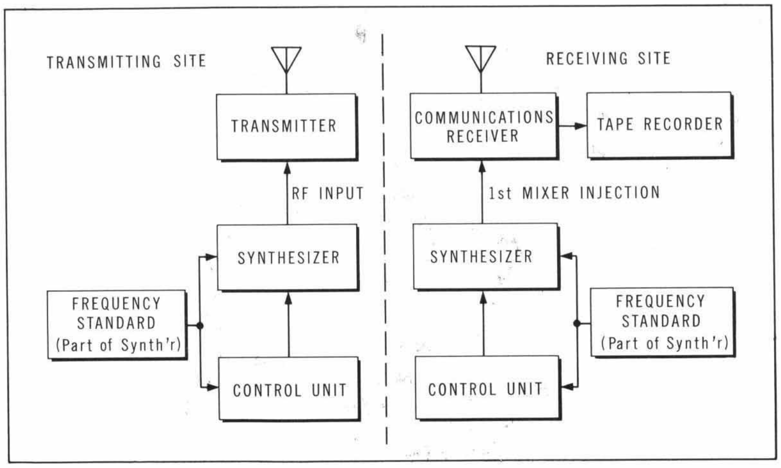


Fig. 5. *Block diagram of FM radar system used in experiments described in text. Receiver system is located some distance from transmitter site to reduce direct signal between sites and uses second synthesizer to translate received echo to IF of receiver. Second synthesizer tracks transmitting synthesizer with appropriate frequency offset.*

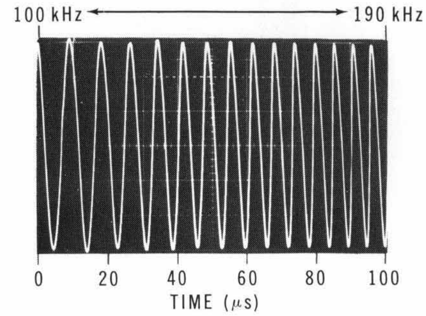


Fig. 7. *Oscillogram shows example of swept frequency output of modified synthesizer.*

available). Seen through a receiver bandwidth of a few hundred cycles, the hf spectrum is nearly unoccupied. Further, the receiver may be gated off when unusable frequencies *are* encountered, removing the interference with negligible loss of desired signal. The analogous process may, of course, be performed by inserting multiple notches in the frequency response of any sounder but the rejection is notably simpler to implement in the time domain.

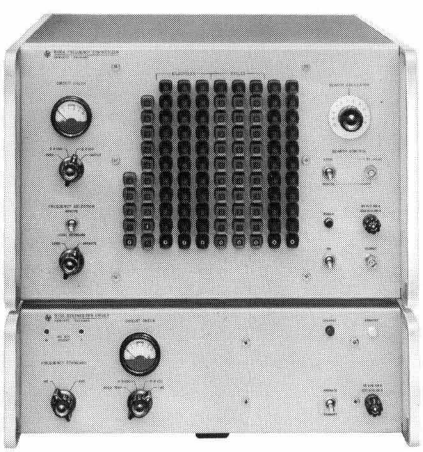


Fig. 6. *–hp– Model 5100/5110 Frequency Synthesizer generates high-stability signals in 0.01 Hz increments from dc to 50 MHz. Output frequency can be controlled manually by panel keyboard or remotely by electrical signals. For frequency sweep applications, synthesizer is modified, with some sacrifice in spectral purity, to permit frequency change in 0.1 μs; switch is added to permit phase alignment of internal signals at turn-on by manual interruption of divider circuits.*

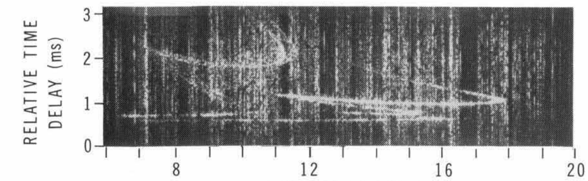
Another advantage of FM systems is the simplicity of data reduction possible with a linear sweep. In order to generate a display of echoes versus range, more general pulse-compression techniques require either a tapped delay line having an accuracy commensurate with the system time-bandwidth product, or a separate cross-correlation analysis for every range interval of interest. A range display can be obtained from a Chirp radar simply by spectrum-analyzing the receiver output. This convenience arises as a result of the linear frequency sweep. A fixed time delay Δt corresponds to a fixed-frequency offset Δf, where $\Delta f = \frac{df}{dt}\Delta t$. The spectrum (amplitude-frequency) of the received signal thus becomes the normal A-scope (amplitude-range) display.

SYNTHESIZING A LINEAR FREQUENCY-SWEEP

The linear-sweep radar devised at Stanford is shown in the basic block diagram of Fig. 5. This system makes use of the fact that a variable-frequency signal may be approximated by a succession of fixed-frequency segments to whatever accuracy may be desired (Fig. 10). A linear frequency-sweep is particularly simple to achieve since the successive frequencies are equally spaced and are used for equal time intervals. The –hp– Model 5100A/5110A Frequency Synthesizer provides a choice from among five billion fixed frequencies, all equally spaced and directly synthesized from a frequency standard. It is tempting then to hope that a nearly-ideal linear sweep might be generated by simply controlling the frequency of the synthesizer from an instrument similar to a digital time code generator. This is essentially the case.

If a piecewise constant-frequency is to approximate a continuously-varying one, the transitions from one frequency to the next must be made with phase continuity. Otherwise, the result is a phase-coded pulse rather than a frequency sweep. Phase continuity can be obtained at the output of the synthesizer by insuring that all frequency switching within the synthesizer is performed at instants when the "old" and "new" frequencies have identical phase. Fortunately, all frequencies produced by the –hp– 5100A Synthesizer are selected by switching from among signals spaced at multiples of 100 kHz, so instants of phase equality between new and old signals can be relied upon to occur every 10 μs. If, on the other hand, the output were chosen from a large number of separate oscillators, the problem would be much more difficult. A frequency spacing of, say, 1 Hz between adjacent oscillators would result in the occurrence of phase equality only once per second. The approximation to a continuous frequency sweep which could be generated with such infrequent switching would be poor indeed.

Modification of the synthesizer is necessary to insure that *all* possible signal components within the synthesizer arrive at the desired switching



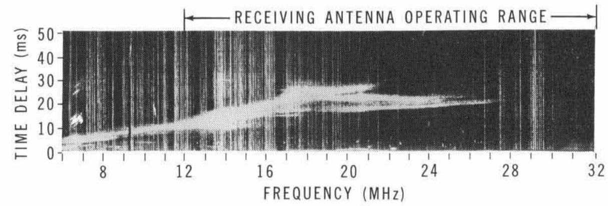


Fig. 8. *Ionograms made by oblique sounding over 2000-km path with 20-mW Chirp radar (left) and by backscatter technique with 100-W Chirp radar (right). Lowest line (left) is signal reflected by sporadic E layer; upper curved lines are (in order) E-layer and one- and two-hop F-layer signals. Diffuse nature of return signal (right) occurs as result of groundscatter from range continuum.*

phase together. When this is done, a change can be made to any desired new frequency, without cumulative phase errors, at instants separated by 10 μs or any multiple of 10 μs.

Switching the synthesizer every 10 μs, however, requires modifications of internal frequency gating circuits. The frequency-switching speed of the -hp- synthesizer is only specified as less than 1 ms, although most frequency changes are completed in much less than this time. Switching in the modified synthesizer is completed in the order of 100 nanoseconds.* A photograph illustrating frequency sweep from 100 kHz to 190 kHz in 100 μs is shown in Fig. 7.

EXPERIMENTAL RECORDS

The reality of Chirp radar's combined advantages of high average-to-peak power ratio and interference rejection is demonstrated by the comparison of Fig. 2. The lower record was taken at Stanford with a conventional Ionosonde (Model C-2). The sounder transmits 50-μs pulses at a peak power of 2 kW. The upper record was made using the Chirp radar with a modest output power (actually, the 20 milliwatts from the synthesizer applied directly to the antenna). The Chirp echoes were recorded on magnetic tape and later spectrum-analyzed to make the record shown. Since the recorded signals represent a continuous sweep across the entire hf spectrum, the tape can be analyzed using various filter bandwidths to provide any desired trade-off between frequency resolution and echo-time-delay resolution consistent with the dispersion and delay of ionospherically-reflected signals. A time resolution of 20 μs is generally a good compromise for ionospheric sounding, and a spectrum analyzer bandwidth equivalent to 20 μs was used in preparing Fig. 2a.

The 20-mW output from the synthesizer is also adequate to make acceptable oblique ionograms as shown in Fig. 8. A more demanding application is the case of backscatter sounding, where the received power is a much smaller fraction of that transmitted. Prior to the advent of sweep-frequency sounding, the available transmitter power has been taxed to produce records of the desired resolution, especially in the presence of the inevitable interference. A 100-W FM sounder has been shown to have resolution comparable to a conventional pulse-sounder operating with 30-kW, 100-μs pulses (Fig. 8).

In lunar radar studies conducted by the Stanford Center for Radar Astronomy, ease of data reduction provided the motive for use of the Chirp technique. Experiments are in progress to explore the scattering behavior of the lunar surface and to determine the electron content between the earth and the moon. Both demand a radar range resolution of 10 μs or better. While coded pulse transmissions were originally employed for the lunar sounding, the data-handling problems were severe, requiring analog-to-digital conversion and recording of several channels of wideband signals for later cross-correlation on a computer. The Chirp echo signals occupy only a narrow bandwidth and are easily analyzed with an audio-frequency analyzer. The system is employed routinely to obtain records like that shown in Fig. 1.

SWEEP ACCURACY

It is interesting to consider the degree to which a synthesized frequency sweep departs from the ideal linear sweep. Ignoring inaccuracies in phase alignment within the synthesizer, consider the spurious signals which arise from the staircase frequency characteristic of the synthesized signal. The solid curve of Fig. 9a represents the synthesized sweep and the dotted curve is the desired ideal linear sweep. The difference between them is the sawtooth frequency error of Fig. 9b. Integrating the frequency error gives the phase error, a series of parabolic curves shown in Fig. 9c. The peak-to-peak phase error θ_e may be easily shown to be $\frac{\pi}{4} f_i t_i$ radians, where f_i and t_i are the frequency and time increments of the synthesizing sweep.

The importance of this phase error is most easily seen by imagining that the ideal linear sweep is represented by a phasor (Fig. 10a). The actual synthesized phasor oscillates about the ideal phasor as shown in Fig. 10b. For small p-p phase error, the oscillating phasor may be considered as the sum

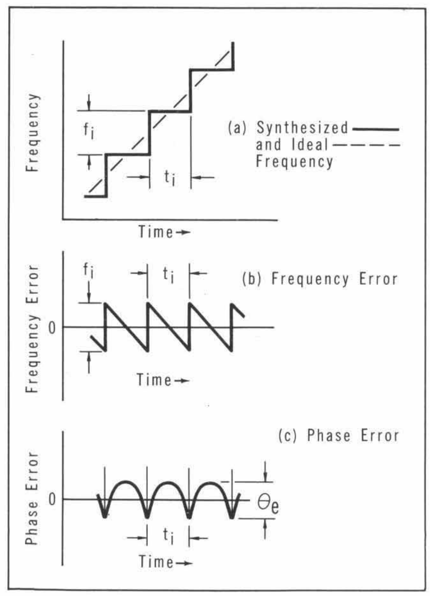


Fig. 9. *Synthesized frequency-sweep is shown in (a) as step-wise approximation to ideal sweep. Resulting frequency and phase errors are shown in (b) and (c).*

* Modified synthesizers together with appropriate control circuitry are now available commercially from Applied Technology Incorporated, Palo Alto, California.

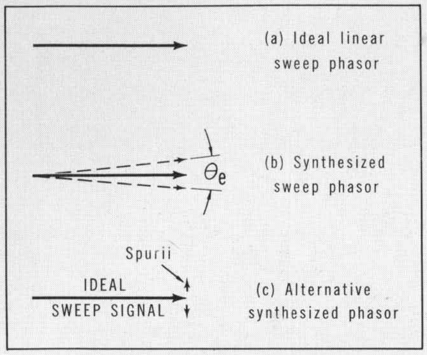


Fig. 10. *Representation of frequency sweep by phasors. Phasor representing synthesized sweep (b) oscillates back and forth through peak-to-peak phase error θ_e and may be represented by vector addition of ideal sweep phasor and small amplitude-modulated signal in phase quadrature (c).*

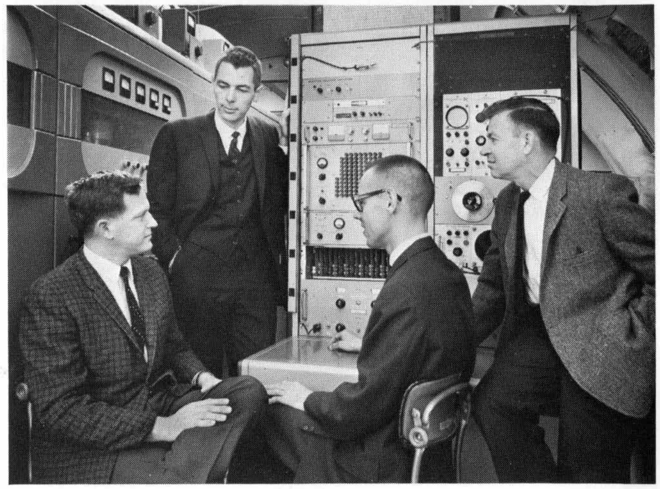


Fig. 11. *Drs. George Barry (standing left) and Robert Fenwick (center) of Stanford University's Radioscience Laboratory discuss synthesized frequency sweep system with Victor Van Duzer (seated left) and Alex Tykulsky of the -hp- Frequency Synthesizer development group.*

of the ideal phasor and an amplitude-modulated spurious signal in phase quadrature (Fig. 10c). The resulting sidebands may be readily calculated in detail and the results show the total power in these spurious sidebands to be lower than the desired signal by an amount

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

Most synthesized Chirp sounding to date has employed a sweep rate of 100 kHz/sec or frequency and time increments (f_i and t_i) of 1 Hz and 10 μs, respectively. These values give a total spurious power of —111.4 dB relative to the ideal sweep signal. Obviously, for these values, the spurious output level is determined by equipment imperfections rather than the stepwise frequency approximation. Even for sweep rates as high as 100 MHz/sec, the approximation results in spurii over 50 dB below the ideal signal.

Generally, a much more important source of phase errors is the residual phase misalignment which exists at times of frequency switching. The spurious signal components which result from these phasing errors lie close to the ideal signal, rather than 100 kHz or further away as is the case of energy from staircase-frequency errors.

FUTURE APPLICATIONS

The potential applications for precise variable-frequency waveforms appear to be limited only by one's imagination. In principle, *any* waveform can be approximated to arbitrary accuracy by the technique. Immediately obvious applications are further uses of linear sweeps for sounding — throughout the radio spectrum and for sonar and geological exploration as well.

Non-linear synthesized sweeps have not yet been put to use. Possible applications occur in the fields of network measurement, communications and Doppler tracking.

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1. J. R. Klauder, A. C. Price, S. Darlington, and W. J. Albershein, "Theory and Design of Chirp Radars," **BSTJ,** 39, July 1960, pp. 745–808.
2. Kenneth Davies, **Ionospheric Radio Propagation,** NBS Monograph No. 80, National Bureau of Standards, Boulder, Colorado, April 1965.

VOLTAGE AND TDR MEASUREMENTS TO BE DISCUSSED AT WESCON/65 TECHNICAL SESSION

Voltage Measurements, from DC to Microwave, and Time Domain Reflectometry will be discussed in depth during Technical Session No. 8 at WESCON/65 (Aug. 24-27). This is one of the WESCON technical sessions organized by industry, each of which will present a "project team" discussing a single project or program in closely related papers. Session 8, organized by Hewlett-Packard, is devoted to electronic measurements covering the two topics: voltage measurements and TDR. The program, to start at 10:00 a.m. on Aug. 25, is scheduled as follows:

Transmission Line Measurements in the Time Domain
J. Sedlmeyer, Edgerton, Germeshausen, and Grier, Las Vegas

Time Domain Reflectometry as a Design Tool
Carl Sontheimer, Anzac Electronics, Norwalk

Scaling of Microwave Components
H. Poulter and S. Adams, Hewlett-Packard, Palo Alto

Voltmeter Calibration to 1 GHz
Myron C. Selby, National Bureau of Standards, Boulder

Coaxial Line Standards for Measurement of Reflection with a Time Domain Reflectometer System
Jose Cruz, National Bureau of Standards, Boulder

DC Voltmeters — Accuracy Specification, Traceability, and Verification
Donald F. Schulz, Hewlett-Packard, Loveland

Extraneous Signals which make DC Voltage Measurements Difficult
Paul G. Baird, Hewlett-Packard, Loveland

Measurement Errors Introduced by Distorted and Non-Sinusoidal Signals when Detected with Peak, Average, and RMS Responding Voltmeters
Marco R. Negrete, Hewlett-Packard, Loveland

Co-chairmen of the session will be Marco Negrete, Engineering Manager of the -hp- Loveland Division, and Darwin Howard, Engineering Manager of the -hp- Oscilloscope Division.