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The AR6A Single-Sideband Microwave Radio System:

Microwave Carrier Supply

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In the AR6A repeater bay, microwave carrier power for both the receiver and the transmitter is derived from a single voltage-controlled crystal oscillator. For long-term stability, this oscillator is phase locked to an external-reference frequency, available at each radio station. From the output signal of this oscillator, an active frequency-multiplier chain generates about +21 dBm power in the 6-GHz band. Part of this power serves as the local oscillator signal for the transmitter directly. The other part is shifted in frequency, to produce the local oscillator signal for the receiver.

I. INTRODUCTION

Two microwave carriers are needed in an AR6A† repeater bay for up and down conversion. These carriers have to be spaced 252 MHz apart and must have low noise, low jitter, and exceptional frequency stability.

In the microwave carrier supply, both carriers are derived from a single Voltage-Controlled Crystal Oscillator (VCXO)‡. An active frequency-multiplier chain generates the 6-GHz carrier for the up con-

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† Amplitude Modulation Radio at 6 GHz for the initial (A) version of the system.

‡ Acronyms and abbreviations used in the text and figures of this paper are defined at the back of this *Journal.*

verter directly. Part of this 6-GHz power is split off and shifted 252 MHz in frequency, to be used in the receiver down converter.

For frequency stability, the VCXO is phase locked to an external reference, available at each radio station. Steps are taken to ensure adequate frequency stability even during temporary loss of the external-reference signal.

Along with the detailed performance objectives, a general circuit description of the microwave generator is given in this paper, with the key circuits described in somewhat more detail. Typical test results and a brief description of the method used to measure FM noise close to the carrier concludes the paper.

II. PERFORMANCE OBJECTIVES

Some of the performance objectives for the microwave carrier source designed for AR6A are similar to those required in FM radio systems; others are unique. The performance objectives for this new carrier source are enumerated below with discussion of those items dictated by application in the AR6A Radio System.

2.1 Frequencies

The TR-bay microwave carrier supply has to provide two carriers separated by 252 MHz, in the 6-GHz band at any of the 10 Local Oscillator (LO) frequencies given in the AR6A frequency plan.[1]

2.2 Power output

The available microwave carrier power requirement is +17 dBm for the transmitter and +11 dBm for the receiver at the shifted LO frequency.

2.3 Frequency stability

The microwave carriers must be stable in frequency to about two parts in 10^7 per year so that the equalizer pilots at IF will be located accurately with respect to the narrowband (12 kHz) crystal pick-off filters in each AR6A receiver. In addition, the accumulated frequency error from each repeater in a terminal section must be limited to a value that can be corrected by the multimastergroup terminal receiver. This stability objective of two parts in 10^7 is about two orders of magnitude more accurate than for carriers for FM radio application. To meet the strict frequency-stability requirement, the microwave carriers are phase locked to an external-reference frequency,[2] having the required frequency accuracy and stability.

With conventional phase-lock control, if lock is lost the error-correction signal to the oscillator will go to zero and the frequency of oscillation will shift to a free-running condition that will depend on temperature and oscillator aging since last tuning. In many cases this shift would be sufficient to render the radio channel unusable. To prevent this almost certain channel loss, an additional objective is specified for the phase-locked circuitry that on loss of lock the oscillator control voltage be maintained to keep the 6-GHz output frequency within ±200 Hz of the value when loss of lock occurred. Under these conditions the microwave generator free runs and drifts with temperature and time from this frequency.

Large temperature changes in a nonair-conditioned repeater station during a time that the microwave generator is free running could cause excessive frequency drift. Such large changes may occur from day-to-night temperature variations. To minimize these effects, a final objective on frequency stability is that the free-running frequency shall not change by more than 0.5 ppm over any 30-degrees F ambient temperature change within the limits of 40 to 140 degrees F. This objective is intended to provide acceptable frequency stability during the time the source is free running over a period of days if technicians cannot be immediately dispatched to clear the loss-of-lock condition.

The free-running frequency stability, together with the phase-locked error voltage memory, ensures continuous service on the AR6A route even in the unlikely event of the loss of the reference signal from the Microwave Carrier Synchronization Supply (MCSS), when every microwave source in the station is unlocked.

2.4 Phase noise

The phase noise requirements for a microwave source used on an AM radio system are more stringent than for use on FM systems. For an AM system the microwave source phase-noise spectrum close to the carrier is dominant in the contribution of voice-circuit noise.

To demonstrate this fact, consider a baseband signal $a(t)$ that is up converted in frequency by a local oscillator with frequency ω_c and phase noise $\theta(t)$. If transmitted single sideband, the signal $s(t)$ can be represented as

$$s(t) = a(t)\cos[\omega_c t + \theta(t)] \pm \hat{a}(t)\sin[\omega_c t + \theta(t)], \tag{1}$$

where $\hat{a}(t)$ is the Hilbert transform of $a(t)$. The sign of the second term is negative if the upper sideband is selected and positive if lower sideband is selected. To evaluate the noise contribution from one source, assume the demodulation is done by a noise-free source. The applicable output terms, $r(t)$, at baseband are given by

$$r(t) = a(t)\cos[\theta(t)] \pm \hat{a}(t)\sin[\theta(t)]. \tag{2}$$

For the LO source $|\theta(t)| \ll 1$, so that $\cos[\theta(t)] \approx 1$ and $\sin[\theta(t)] \approx$

$\theta(t)$. In this case eq. (2) becomes

$$r(t) \approx a(t) \pm \hat{a}(t)\theta(t). \tag{3}$$

The output consists of the desired signal $a(t)$ plus a noise term given by $\hat{a}(t)\theta(t)$. Let $A(f)$ denote the power spectral density of the desired signal and $S_a(f)$ the spectral density of the phase noise. The spectral density of the Hilbert transform of a signal is equal to the spectral density of the signal itself; therefore, the spectral density, $N(f)$, of the noise term in eq. (3) is

$$N(f) = A(f)*S_a(f), \tag{4}$$

where $*$ denotes convolution.

The spectral density of the 6000 voice-circuit channel is noiselike and essentially flat over the 30-MHz frequency band. Since the phase spectral density of crystal oscillators as used in the microwave source increases as $1/f^3$ close to the carrier, the close-in noise components contribute the most to the convolution.

Since the phase jitter is present on the carrier recovery pilot, the Multimastergroup Translator for Radio (MMGT-R)[3] tracking receiver at the end of a terminal section will remove some of this noise. The remaining phase noise will not be reduced further by the following section since new carrier recovery pilots are inserted at the beginning of the next span. The total phase noise will accumulate as each terminal section is traversed. Taking into account the phase-noise reduction that can be obtained from the terminal, the phase-noise spectral density objective of the microwave generator necessary to meet its 4000-mile noise allocation was determined. This objective is shown as the upper curve in Fig. 1.

Though not part of the dBrnc0 noise-tree allocation, the phase noise of the microwave sources will affect voice-circuit phase-jitter objectives, which are important when the channels are used for voiceband data transmission.

III. MICROWAVE-GENERATOR CIRCUIT DESCRIPTION

Figure 2 shows the block diagram of the microwave carrier supply for an AR6A repeater. A 21.5-dBm, 6-GHz signal is generated using a 1-GHz microwave generator and a 6X frequency multiplier. The 6-GHz power is split in the carrier distribution network, with +17 dBm going directly to the transmitter modulator; the remainder drives the shift modulator to generate the carrier for the receiver modulator. The shift oscillator frequency is 252 MHz.

Both the 1-GHz generator and the shift oscillator are phase-locked to a central 308.9-kHz reference signal through their respective frequency control units. In case of a reference-signal failure, the frequency

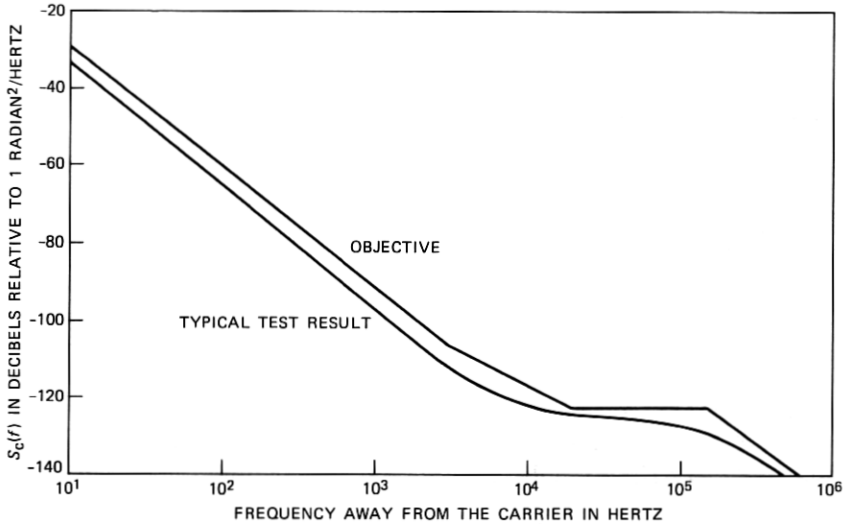

Fig. 1—Phase-noise objective and typical test result for the microwave generator.

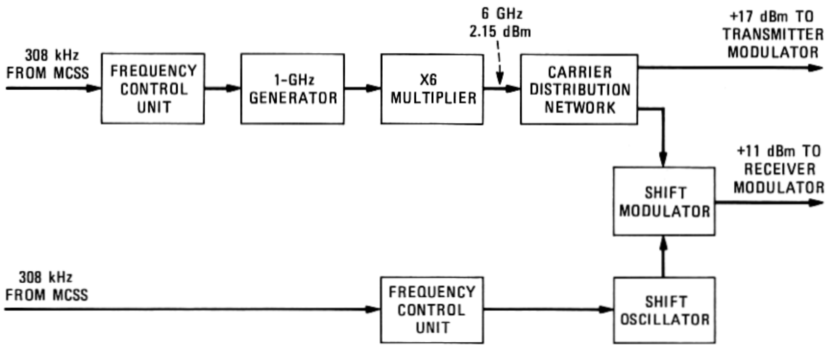

Fig. 2—Microwave carrier supply.

control units will put the oscillators on memory in order to continue to supply the microwave carrier and keep their frequencies within the required accuracy.

3.1 1-GHz generator

At the beginning of the AR6A development, several candidates had been investigated for the 1-GHz generator, including general trade products. For stability, performance, and cost effectiveness the 1-GHz generator developed earlier for the TH-3 Radio System[4] was adopted and improved for AR6A radio use. Figure 3 shows the block diagram of the 1-GHz generator.

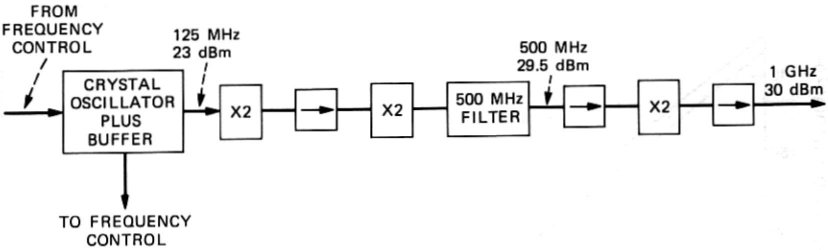

Fig. 3—1-GHz generator.

The generator is a straight multiplier chain of three active frequency doublers, driven by a crystal oscillator in the 125-MHz frequency band. Three isolators are used between successive stages to stabilize the doublers and facilitate tuning. A high Q INVAR cavity is used at 500 MHz to strip off multiplied white noise far from the carrier.

3.1.1 Oscillator and buffer amplifier

The crystal oscillator is built with a Western Electric NPN transistor in the common base configuration. The third-overtone crystal operates in series resonance, and it is connected in the feedback loop between the emitter and collector of the transistor. A high Q varactor diode is connected in series with the crystal to provide voltage control of the oscillator frequency.

To meet the AR6A requirement for low noise close to the carrier, it was important to keep the $1/f$ noise of the transistor low by keeping the collector current low, to use a crystal with high unloaded Q, and to design the oscillator circuit (including the varactor) such that the Q of the total oscillator would approach the Q of the crystal itself. To keep intermodulation in the transistor at minimum, and to keep the conversion of $1/f$ noise into the 125-MHz frequency band low, limiter diodes are used to control the amplitude of oscillation, instead of relying on the nonlinearities of the transistor itself.

For long-term stability the crystal current is kept below 1 mA, which gives about +7 dBm power at the oscillator output.

Under normal operating conditions, the varactor voltage is regulated by the Phase-Locked Loop (PLL) in the frequency control unit to keep the crystal-oscillator frequency locked to a high-order harmonic (404th to 429th, depending on channel frequency) of the 308.9-kHz reference signal. In case of loss of lock (because of an MCSS failure, for example), the frequency control unit goes on memory, and the varactor voltage in the crystal oscillator is kept constant at the last (locked) value. In this free-running mode the crystal oscillator has to stay within 1 ppm of its nominal frequency until the unit is repaired (two days at most). To meet this requirement in stations without air conditioning, the crystal is placed in an oven. The turnover temperature of the crystal is 65 degrees C.

A two-stage wideband buffer amplifier follows the crystal oscillator. A shunt-mounted PIN diode between the two stages of the buffer can be used to adjust levels for the entire microwave generator. The output level range of the buffer is 13 to 25 dBm.

A sample of about 12 dBm is decoupled from the buffer output to the frequency control unit for phase locking.

3.1.2 Frequency doublers

The three active frequency doublers in the 1-GHz generator are built with Western Electric overlay transistors. Besides doubling their respective input frequencies, these stages have several decibels of gain each. Figure 4 shows a common simplified schematic of the doubler circuits.

Due to the ac short in the collector circuit at the input frequency, *f*, these stages have substantial current gain at that frequency. The increased collector current pumps the nonlinear capacitance of the collector-base junction, producing harmonic voltage components between collector and base. A second series resonant circuit between base and ground provides a current path for the generated second harmonic, bypassing both the input matching network and the base emitter junction of the transistor. The two matching networks transform the input and output impedance of the doubler circuit to 50 ohms.

Principally, all three doubler stages are as described. Only the impedance levels vary with frequency and the physical form of the circuit elements. In the first doubler from 125 to 250 MHz most elements are lumped; the third stage, however, from 500 MHz to 1 GHz is built entirely of distributed elements.

Nominal output power levels for the three doublers are: 27 dBm at 250 MHz, 30 dBm at 500 MHz, and 31 dBm at 1 GHz. With interstage losses taken into account, this corresponds to conversion gains of 4, 3, and 2 dB, respectively.

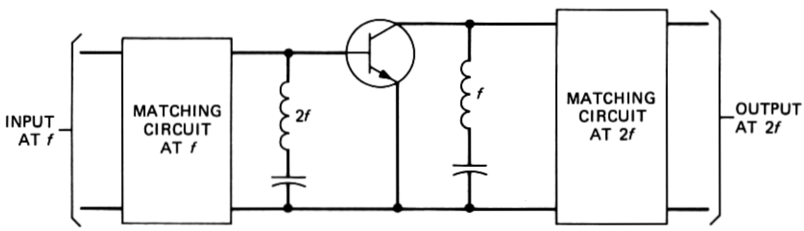

Fig. 4—Transistor doubler.

3.1.3 Isolators

All three isolators in the multiplier chain are of the lumped element design.[5] They have been developed at Bell Laboratories, Allentown, Pennsylvania. Their insertion loss is typically 0.5 dB forward, and greater than 20 dB in the reverse direction.

3.1.4 Noise-suppression filter

Although the excess noise of the frequency doublers is minimal, they increase the FM noise originating in the oscillator and buffer by a factor of 20 log n along the multiplier chain. Far from the carrier, therefore, the multiplied noise could exceed system requirements.

To prevent this, a noise-suppression filter is inserted into the multiplier chain at 500 MHz.[6] The filter is a coaxial reentrant cavity with an unloaded Q of about 3000. Its 3-dB bandwidth is about 300 kHz, with about 2-dB insertion loss. For frequency stability the cavity is extruded from INVAR.

3.2 The 1- to 6-GHz multiplier

The 1- to 6-GHz multiplier is a broadband 6X frequency multiplier,[7] driven by the 1-GHz generator at a 29.5-dBm level. It delivers 21.5-dBm, 6-GHz power to the carrier distribution network. All unwanted harmonics are kept at least 80 dB below the carrier.

The multiplier circuit is built on alumina microstrip. The photograph in Fig. 5 shows the circuit details. Figure 6 gives the approximate lumped element equivalent circuit.

A self-biased commercial step recovery diode with zero bias capacitance of 3.3 pF is used for the nonlinear element. The diode is mounted into a hole in the microstrip circuit. One side is soldered to the ground plane on the back, the other side is thermocompression bonded to the microstrip pattern using a 20-mil wide gold ribbon.

Both the input and output filters are three-resonator bandpass, 0.1-dB Tchebyscheff type, designed to transform the diode impedance to 50 ohms in their respective passbands. The input filter is realized using quarter-wavelength resonators in an interdigital configuration. The output filter consists of side-coupled, half-wavelength resonators. The diode itself is part of the first resonator in the output filter. The radial lines at the input side of the diode are used to suppress unwanted harmonics; at the same time they act together as a capacitor for the pulse-forming circuit.

In order to meet the very strict spectral purity requirement for the AR6A System at the up and down converter inputs, as discussed earlier, all the unwanted harmonics had to be suppressed to at least 80 dB below the carrier at the 6X multiplier output. This requirement was met by adding a seven-resonator interdigital filter to the output

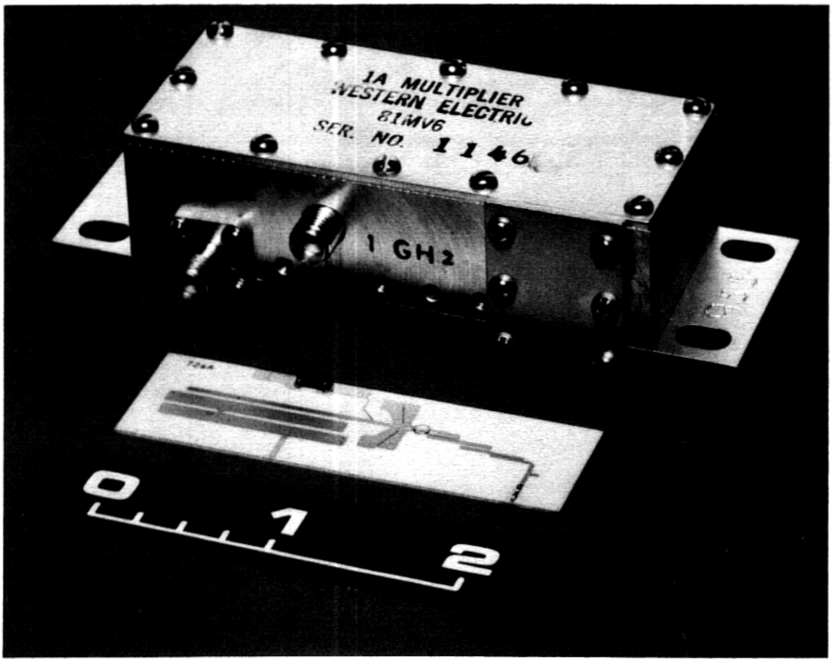

Fig. 5—Photograph of the 6X multiplier.

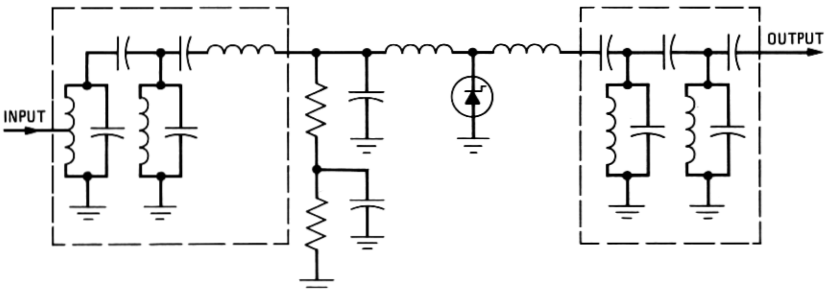

Fig. 6—Lumped element equivalent circuit of the 6X multiplier.

of the 6X multiplier. This filter is built into the back of the multiplier housing, and it has about 0.5-dB passband insertion loss. The overall size of the 6X multiplier, excluding the SMA connectors, is 3/4 by 1/4 by 3 inches. The typical swept response of the multiplier is shown in Fig. 7. One code covers the total AR6A bandwidth.

3.3 Carrier distribution network

As indicated in Fig. 2, the carrier distribution network is used to split the output signal of the microwave generator between the trans-

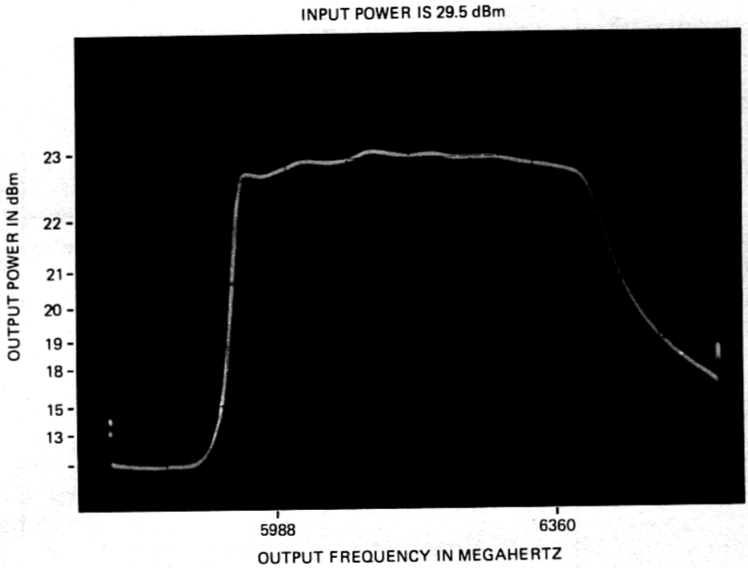

Fig. 7—Typical swept response of the 6X multiplier.

mitter and the shift modulator in a repeater station. The detailed description of this circuit can be found elsewhere.[8]

3.4 *Shift modulator and shift oscillator*

The purpose of the shift modulator is to shift the frequency of the microwave generator by ±252 MHz and thus provide the local oscillator signal for the receiver modulator.

The shift oscillator generates the 252-MHz signal for the shift modulator using a 126-MHz crystal oscillator and a frequency-doubler circuit. As shown in Fig. 2, the shift oscillator is phase locked to the MCSS signal.

Reference 9 gives a more detailed description of both the shift modulator and the shift oscillator.

IV. FREQUENCY CONTROL UNIT

The microwave generator and shift oscillator are phase locked to a harmonic of the microwave carrier synchronization supply frequency of 308.9 kHz† by means of two frequency control units. These units contain a comb generator to produce the desired harmonic of the 308.9-kHz signal; a phase-locked loop to synchronize the microwave generator or shift oscillator to the proper 308.9-kHz harmonic; a memory circuit that operates in conjunction with the phase-locked

† The actual frequency is 308.8735416 kHz.

loop to provide the control voltage to the oscillators; and alarm circuitry to provide bay alarms on loss of lock and memory end of range. The overall description of operation and some details of these individual circuits are given in the following sections.

4.1 Frequency control unit circuit description

Figure 8 shows the block diagram of the frequency control unit. The 308.9-kHz reference signal from the MCSS is applied to the comb generator. The proper harmonic for locking the crystal oscillator is produced by the comb generator and applied to one input of the phase detector in the phase-locked circuit. A signal sample of the crystal oscillator to be controlled is applied to the other phase-detector input of the phase-locked loop. The output of the phase-locked loop is an analog error signal proportional to the phase difference between the reference and controlled oscillators. This error signal is applied as one input to the varactor of the controlled oscillator, with the other inputs from the memory circuit and a bias voltage source. The purpose of the fixed bias is to place operation of the varactor diode at the suitable point on its voltage-versus-capacitance characteristic and to offset the midrange digital error-correction voltage to zero. On initial alignment no digital error-voltage component (after offset) is applied to the varactor diode. In time, due to aging and temperature changes, the analog error voltage will change to maintain lock, and when the change reaches a certain value, the memory circuit will add a digital error-voltage component of a value equal to the change. In order to maintain the total error signal the same, the analog component will return back to its original value. As further changes occur, the memory circuit will continue to add or subtract steps as the analog signal either increases

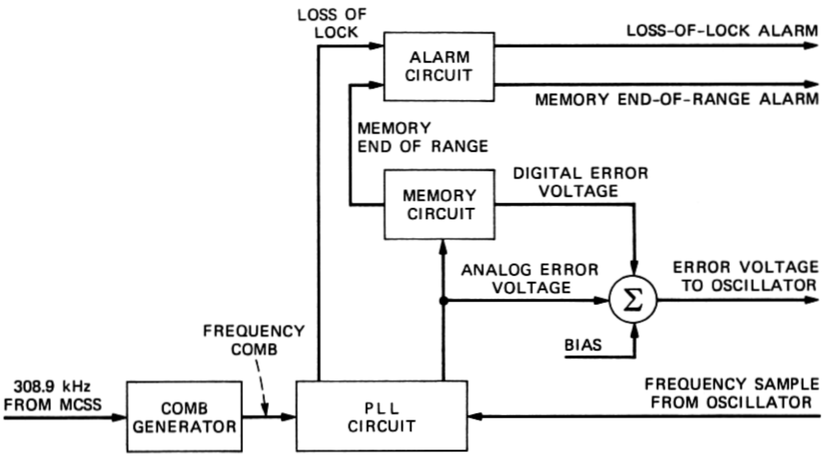

Fig. 8—Block diagram of frequency control unit.

or decreases. In this way the memory circuit provides the major part of the error signal with the phase-locked loop error signal varying an amount equal to the memory change threshold. If the reference signal from the MCSS is lost, the analog loop is opened and the analog error signal goes to zero. The memory continues to supply the digital error-signal component. In this way the generator is free running very near its last correct frequency. The maximum error is plus or minus the memory threshold value which is about 3.2 Hz at 125 MHz, which translates to a maximum error of about 155 Hz at 6 GHz.

There are 1024 steps of memory available with step 512 initially set when the oscillator is aligned. This provides for 512 steps in either direction to compensate for increase or decrease in frequency. This corresponds to ± 1600 Hz at 125 MHz or about ± 1.28 parts in 10^5.

The alarm circuit monitors the memory step location and generates a bay alarm when 88 percent of the available range has been used. This is a warning to the operating personnel that manual retuning of the oscillator should be done. In addition, the alarm circuitry provides an alarm if phase lock between the reference signal and the controlled oscillator is lost.

4.2 Comb generator

The comb generator[10] consists of three basic parts: (1) a sine-wave-to-square-wave converter, (2) a narrow pulse generator, and (3) a gated oscillator tunable in frequency and with output always beginning at the same phase.

The input 308.9-kHz sine-wave reference signal is converted into a TTL-level square wave using a high gain amplifier followed by a comparator. The comparator output is used to trigger a monostable multivibrator that generates 0.5-microsecond pulses at the reference-frequency rate. This narrow pulse is used to gate an oscillator that always starts at the same phase. The frequency of oscillation can be manually tuned. Let f_r denote the frequency of the reference signal, A and f_{osc} the amplitude and frequency of the gated oscillator, and T_w the width of the gating pulse. A Fourier analysis shows that the output spectrum, $S_o(f)$, of the gated oscillator is given by

$$S_o(f) = Af_rT_w \sum_{n=-\infty}^{\infty} \frac{\sin \pi(f_{osc} - nf_r)T_w}{\pi(f_{osc} - nf_r)T_w} \delta(f - nf_r). \tag{5}$$

The output is seen to consist of harmonics of the reference signal with amplitudes proportional to the sin x/x function. By tuning f_{osc} at or near the desired harmonic, say $n_d f_R$, the output of the comb generator will provide maximum output to the PLL at that frequency. Tuning of the oscillator is easily accomplished by observing the beat signal at the output of the PLL phase detector and tuning the oscillator for maximum amplitude. The other harmonics that are present at the phase detector are not important since a given crystal oscillator can never be pulled far enough to lock to the wrong harmonic.

4.3 Phase-lock loop

Figure 9 shows a block diagram of the phase-locked loop circuit. A sample from the controlled oscillator is buffered and amplified by a common base transistor amplifier and applied to one input of the phase detector. The frequency comb is the other input to this detector. The phase detector has a sinusoidal characteristic and is implemented using a double-balanced mixer. In normal operation, a small region around the zero crossing of this characteristic is used to approximate a linear phase detector. The output of the phase detector is amplified to provide the desired PLL loop gain and applied through a Complementary Metal-Oxide Semiconductor (CMOS) Single Pole Double Throw (SPDT) switch to the loop filter and the memory circuit. The output of this filter is the analog portion of the error voltage applied to the varactor.

The CMOS switch is under control of the loss-of-lock detector. As long as the loop is locked the control voltage holds the switch in the through-path position. A prolonged loss of lock will cause the switch to open, which reduces the analog portion of the error signal to zero. At this point only the memory circuit is providing error correction. The reset control forces the switch closed initially so that phase lock can be achieved.

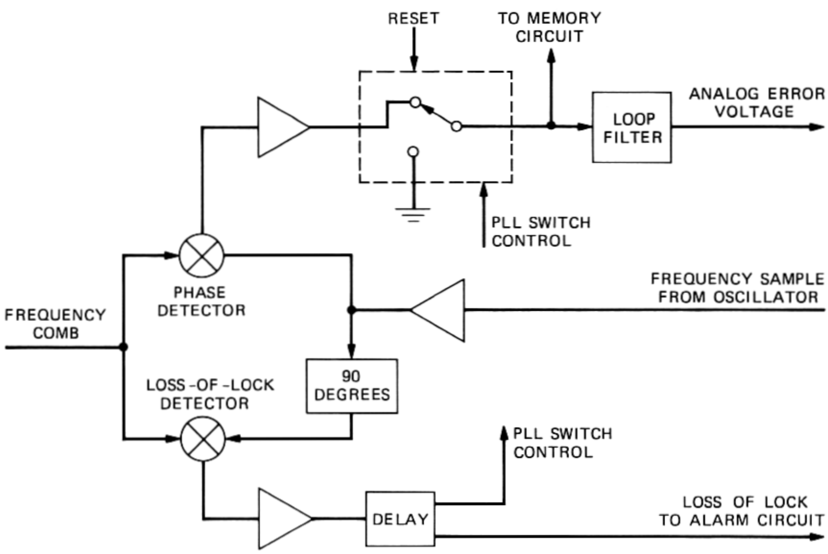

Fig. 9—Frequency control phase-locked loop circuit.

The loss-of-lock detector uses a second double-balanced mixer but with the frequency sample input shifted 90 degrees. When in phase lock, the loop phase detector is operating around the 0-degree point on the characteristic and the lock detector will be operating about the 90-degrees point providing a dc voltage output. This output is amplified and applied to a delay circuit. Loss of lock must occur for the entire period of this delay before the PLL switch is opened and a loss-of-lock alarm given. This delay is necessary to keep the loop closed long enough to reacquire lock lost because of a hit or by a switch in the MCSS from one reference source to another.

Dynamically, the PLL is a second-order type. Since its purpose is for long-term frequency stability, bandwidth is relatively narrow. The primary consideration for bandwidth is that it be adequate to reacquire lock when the MCSS reference switches and that the unit be fairly easy to lock initially when manual tuning of the controlled oscillator is required to bring the frequency within locking range. The pertinent frequency control parameters of the loop are as follows:

1. Natural frequency Ω_n is 51 rad/s.
2. Damping factor is 0.7.
3. Overall loop gain is 2.7×10^3 rad/s.

4.4 Memory circuit

Figure 10 shows a block diagram of the memory circuit. The digital error voltage is derived from a 10-bit Digital-to-Analog (D/A) converter that is driven by a 10-stage up/down counter. On initial alignment the counter is preset to a count of 512, the midpoint of the counter. The output of the D/A converter for this count is offset by the bias source so that the effective digital error signal is zero.

The analog error-voltage sample from the PLL is heavily filtered and fed to two comparators. One comparator is set to switch when the

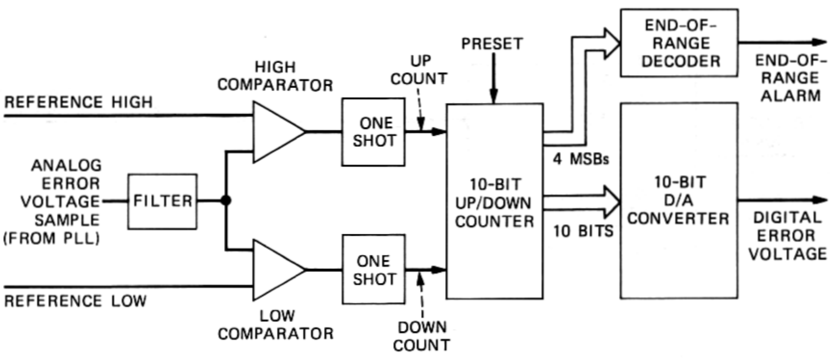

Fig. 10—Frequency control-unit memory circuit.

error voltage indicates a high-frequency correction, the other for a low-frequency correction. The references are set for thresholds corresponding to ±3.2 Hz at 125 MHz. The operation of the comparator activates a one shot, which generates one up-count or down-count pulse to the counter. The scaling of the D/A converter output is such that a 1-bit step will produce a change in output corresponding to a 3.2-Hz correction.

The four most significant bits of the counter are examined for an all 1's or all 0's condition. This indicates that 87.5 percent of the counter's range has been used in the upward or downward direction, respectively. This condition is decoded and output to the alarm circuit as an end-of-range alarm.

4.5 Alarm circuit

The alarm circuit interfaces the frequency control unit to the TR-bay alarm circuits. The end-of-range indication from the memory circuit is latched before it is sent to the TR bay. The loss-of-lock signal is sent directly but buffered by TTL driver gates.

V. PERFORMANCE

5.1 Power output variation with temperature

Figure 11 shows the output power of the microwave generator, measured at the output of the sextupler, as a function of the ambient temperature. The total variation is about 1.6 dB. This is satisfactory for the AR6A System even in a nonair-conditioned environment.

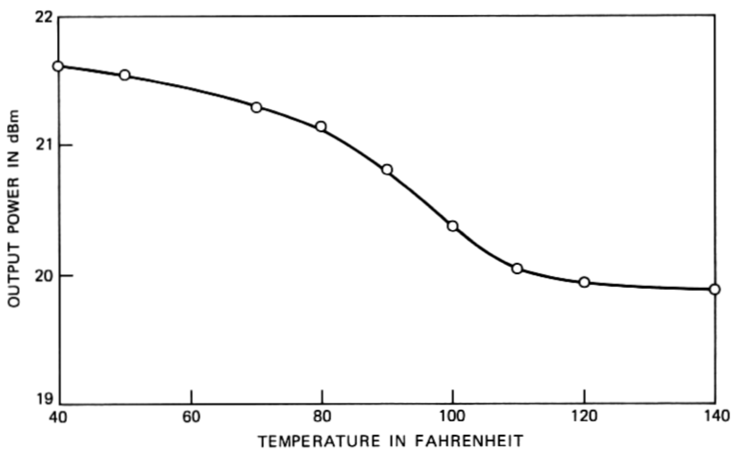

Fig. 11—Output power of the microwave generator as a function of ambient temperature.

5.2 Free-running frequency as a function of temperature

Figure 12 plots the free-running frequency of the microwave generator as a function of the ambient temperature. This characteristic is important only when the oscillator is not locked to the MCSS (due to maintenance or MCSS failure). The variation is within the maximally allowed 0.5 ppm for any 30-degree F change in temperature.

5.3 Phase-noise performance

5.3.1 Measurement method

There are several known methods (Refs. 11–13) to measure phase noise close to the carrier, each of them having its advantages and disadvantages. We have used all of the three methods referred to above during development. Figure 13 shows the block diagram of the phase-noise test set presently used in production. Two microwave generators are phase locked to each other by a narrowband phase-locked loop. The output signal of both are multiplied in frequency up into the 6-GHz band and then fed into a double-balanced mixer in phase quadrature. The output of the mixer goes into a spectrum analyzer after being amplified in a low noise amplifier. The output of the spectrum analyzer is plotted on an X-Y plotter directly as the noise spectral density, S_ϕ(dB) versus frequency separation from the carrier.

5.3.2 Measured phase noise

Figure 1 shows the measured phase-noise spectral density of the microwave generator. The test results are typically 5 dB better than the requirement.

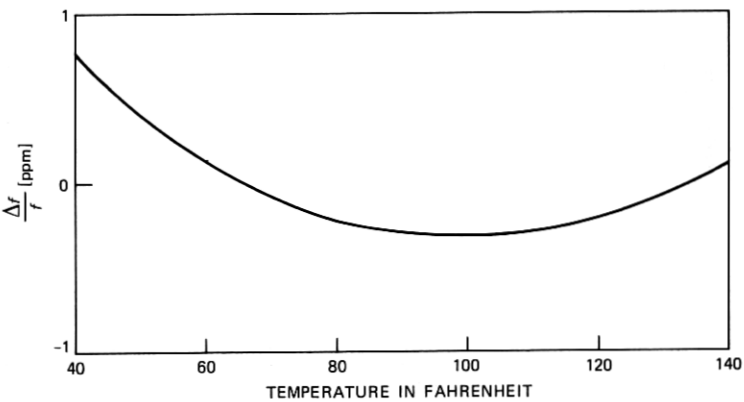

Fig. 12—The measured free-running frequency stability of the microwave generator as a function of the ambient temperature.

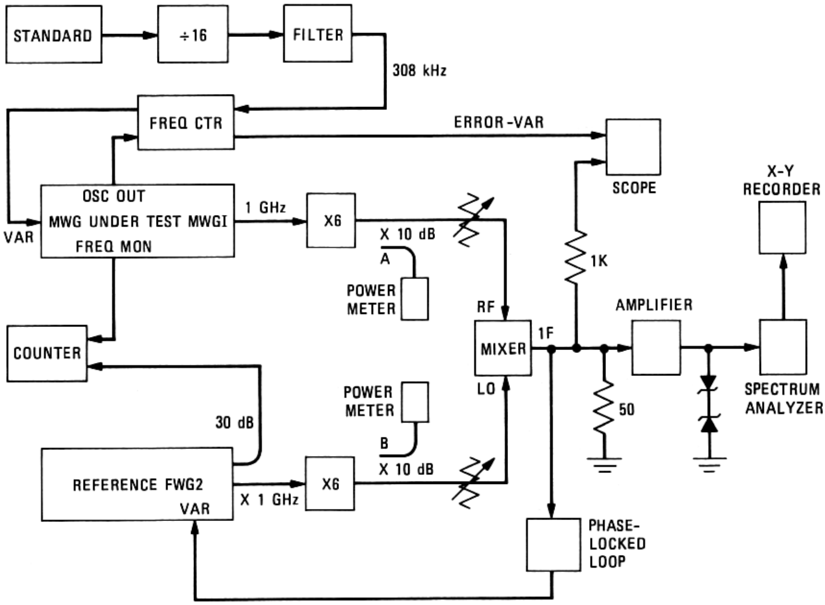

Fig. 13—Block diagram of the test set used to measure noise of the microwave generator.

5.4 Phase jitter

The measured phase jitter of the microwave generator is 10 to 25 degrees peak to peak, using the filter recommended in Ref. 14, and about 0.5 to 0.7 degrees peak to peak with the jitter reduction provided by the MMGT-R and the weighting of the same filter.

5.5 Spurious tones

There are low-level tones in the output spectrum of the microwave generator when the generator is operating in the bay. These tones are harmonics of 60 Hz, coming from the battery power plant, which is charged continuously, and 20- and 40-kHz tones generated by dc-to-dc converters in the power supplies in the bay. All of these tones are at least 40 dB below the carrier at the output of the microwave generator. Also present are 308.9-kHz sidebands from the phase-locked reference. These are better than 70 dB below the desired output.

VI. SUMMARY

A low-noise, phase-locked microwave carrier supply has been developed for the AR6A Radio System with exceptional frequency stability. Steps were taken to ensure system survival even in the extreme case of reference-signal loss.

The architecture and key circuits of the carrier supply have been described, along with performance objectives and typical test results.

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REFERENCES

1. J. Gammie, J. P. Moffatt, R. H. Moseley, and W. A. Robinson, "The AR6A Single-Sideband Microwave Radio System: System Design and Performance," B.S.T.J., this issue.
2. J. M. Kiker, Jr. and S. B. Pirkau, "The AR6A Single-Sideband Microwave Radio System: Frequency Control," B.S.T.J., this issue.
3. A. Dubois, D. N. Ritchie, and F. M. Smith, "The AR6A Single-Sideband Microwave Radio System: Terminal Multiplex Equipment," B.S.T.J., this issue.
4. H. R. Bedell, R. W. Judkins, and R. L. Lahlum, "Microwave Generator," B.S.T.J., *50*, No. 7 (September 1971), pp. 2205–21.
5. R. H. Knerr, "Series and Shunt Tuned Lumped Element Circulators," B.S.T.J., *59*, No. 5 (May-June 1980), pp. 723–30.
6. T. A. Abele and D. J. Leonard, "Microwave Generator," B.S.T.J., *47*, No. 7 (September 1968), pp. 1301–22.
7. J. Gonda, "Broadband Frequency Multiplication by Multitransition Operation of Step Recovery Diode," U. S. Patent 4-360-867, issued November 23, 1982.
8. N. R. Dietrich, "Microwave Integrated Circuits," B.S.T.J., *50*, No. 7 (September 1971), pp. 2175–94.
9. O. Giust, "Modulators," B.S.T.J., *50*, No. 7 (September 1971), pp. 2155–73.
10. M. Emshwiller, "Synchronizing Circuit," U. S. Patent 4-039-968, issued August 2, 1977.
11. J. R. Ashley, C. B. Searles, and F. N. Palka, "The Measurement of Oscillator Noise at Microwave Frequencies," IEEE Trans. Microw. Theory and Tech., *16*, No. 9 (September 1968), pp. 753–60.
12. J. H. Shoaf, D. Halford, and A. S. Risley, "Frequency Stability Specification and Measurement: High Frequency and Microwave Signals," National Bureau of Standards, Technical Note 632, January 1973.
13. L. Peregrino, and D. W. Ricci, "Phase Noise Measurement Using a High Resolution Counter with On-Line Date Processing," 30th Annual Frequency Control Symposium, June 2–4, 1976.
14. "Transmission Parameters Affecting Voiceband Data Transmission Measuring Techniques," Bell System Technical References, PUB 41009, May 1975.

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