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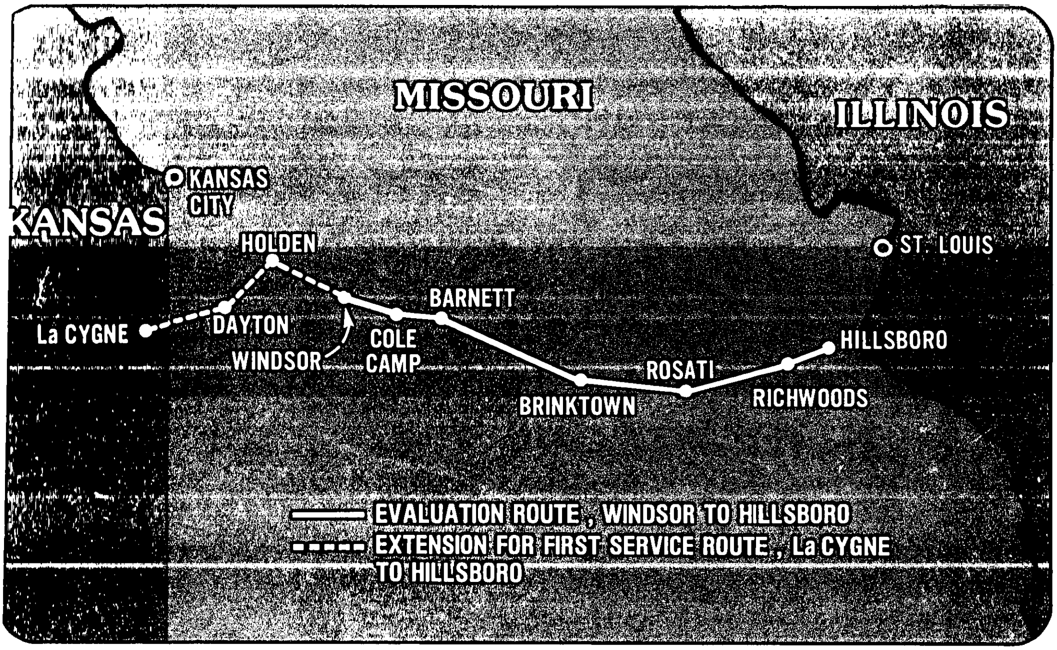
AR6A SINGLE-SIDEBAND MICROWAVE RADIO SYSTEM



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THE AR6A SINGLE-SIDEBAND MICROWAVE RADIO SYSTEM

W. C. Jakes, Jr. *Guest Editor*

THE BELL SYSTEM TECHNICAL JOURNAL
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The AR6A Single-Sideband Microwave Radio System:

Prologue

By R. E. MARKLE*

(Manuscript received January 27, 1983)

The AR6A Single-Sideband Amplitude Modulation (SSBAM) Transmission System makes very efficient use of the radio spectrum by transmitting 6000 message circuits in a 29.65-MHz radio channel. This capability is achieved by maintaining a tighter frequency spacing of the multiplexed mastergroup signals and by discarding redundant sideband signals before transmission over the radio path. The AR6A System can provide a maximum capacity of 42,000 message circuits for the 6-GHz portion of the radio network when all eight channels are utilized. This capability, combined with the alternate use of the 4-GHz spectrum for message or digital circuits, provides the proper mix of network capability choices. This introduction gives an overview of the AR6A System and a brief history of single-sideband evolvement in the Bell System. It also is a prologue to the papers in this volume, which describe some of the specifics of the system design.

I. INTRODUCTION

On January 12, 1981, the first AR6A† Microwave Radio System was placed in service between Hillsboro, Missouri, and La Cygne, Kansas. This event marked the first application of single-sideband transmis-

* Bell Laboratories.

† The AR6A designation represents Amplitude Modulation Radio at 6 GHz for the initial (A) version of the system.

sion to the Bell System long-haul microwave radio network. The AR6A System is designed for use in the 6-GHz common carrier band and can provide a route capacity of 42,000 high-quality, 4-kHz, two-way message circuits when all radio channels in the 6-GHz band have been equipped (i.e., one protection and seven service channels). If TD radio operating in the 4-GHz band is used on the same route at 1800 message circuits per radio channel, the combined route capacity is an impressive 61,800 message circuits.

II. EVOLUTION OF LONG-HAUL RADIO

The evolution of the Bell System long-haul radio network is outlined in Table I; in 31 years the radio channel capacity has been increased by a factor of 12.5 and the route capacity by a factor of 25.75. The ubiquitous TD-2 System was placed in service in 1950 and had a channel capacity of 480 message circuits with a maximum capability of five service channels in the 4-GHz band. In 1954 the circuit loading was increased to 600 message circuits per radio channel, and in 1959 the route capacity was further extended through the use of interstitial radio channels to provide a total of ten service channels. The deployment of TH-1 in the 6-GHz band in 1961 expanded the route capacity by 10,800 circuits to a total of 16,800 message circuits; the advent of

Table I—Evolution of long-haul radio capacities

System/Channel Capacity	Initial Availability	Protection Configuration (Protection × Service)	Route Capacity in Message Circuits
TD/480	1950	1 × 5*	2400
TD/600	1953	1 × 5	3000
TD/600	1959	2 × 10	6000
TD/600		2 × 10	
+	1961		16,800
TH/1800		2 × 6	
TD/1200		2 × 10	
+	1968		22,800
TH/1800		2 × 6	
TD/1500†			
+	1973	2 × 18	28,800
TH/1800†			
TD/1800†,‡			
+	1979	2 × 18	36,000
TH/2400†			
TD/1800†,‡		1 × 11	
+	1981		61,800
AR 6/6000		1 × 7	

* Automatic protection on the basis of one protection channel for five service channels (1 × 5) was not available until 1953; five service channels are used here for consistency in comparing route capacities.
† Each channel also can be equipped with a 1.544-Mb/s system.
‡ The channel alternately can be used to provide 20-Mb/s capability or (beginning in 1983) 45-Mb/s capability.

TD-3 in 1968, with a capacity of 1200 circuits per radio channel, increased the route capacity by an additional 6000 circuits. By 1973, the use of crossband protection switching systems and 1500-circuit loading on TD provided a radio route capacity of 28,800 circuits. Six years later, the circuit loadings of both TD and TH had been increased to 1800 and 2400 circuits, respectively, to give a capability of 36,000 circuits for a fully loaded radio route.

In parallel, the radio systems have grown in digital transmission capability to match network needs. In 1974 data-under-voice capability of 1.544 Mb/s was added to each FM radio channel. The alternate use of a TD-2 radio channel to provide 20-Mb/s capacity was first available in 1979, and an upgrade to 45-Mb/s capacity was provided in 1983.

Even with this capacity some radio routes were expected to be filled by the mid-1980s, and meeting additional growth needs would have required building parallel radio routes or providing some other new facility for relief. It was expected that some of the new radio routes would be difficult to engineer because of radio frequency interference encountered at radio station junctions in the existing radio network. The availability of AR6A in 1981 increased the long-haul radio route capacity to 61,800 circuits and postponed the time when new radio routes would need to be constructed, which in turn postponed the initial capital investment required for land, buildings, towers, and associated station equipment.

III. AR6A SYSTEM FEATURES

The AR6A System allows for orderly growth in the existing radio network, since it uses the same frequency plan as TH-3 and can be used for conversion of TH routes as well as for additions to existing TD routes. The radio equipment is designed to be installed in existing radio stations and shares the use of the radio towers, antennas, common waveguide runs, power plant, and alarm telemetry systems. Additional space-diversity antennas are required over those used on existing FM routes because the single-sideband signal is more susceptible to the effects of multipath fading.

The AR6A System is designed to meet a two-way outage objective of 0.02 percent and a radio channel dynamic-amplitude-misalignment objective of 2 dB or less for 99.9 percent of the time. For a 4000-mile route length, the noise objective is 40 dBrnc0. The system meets these objectives by the appropriate use of space-diversity reception and dynamic equalization in conjunction with a single-channel frequency-diversity protection system.

In meeting intermodulation noise objectives the multiplex, protection, and radio equipment were designed to linearity requirements far more stringent than those of previous radio systems. Of particular note is the AR6A transmitter-receiver (TR)* unit that has a linearity that is more than 30 dB better than the comparable TH-3 TR unit. This is achieved by the use of ultralinear semiconductor devices in the intermediate frequency circuits and predistortion in the transmitter to improve the linearity of the frequency converter and traveling-wave-tube amplifier combination.

The frequency stability required for AR6A is much tighter than that for FM systems (typically by a factor of about 25). This stability allows equalizer pilots to be selected by narrowband crystal filters and reduces the frequency range over which the receiving multiplex terminal must operate.

The AR6A System uses microprocessor-controlled maintenance and surveillance methods to monitor system performance and to isolate troubles to a specific station on a radio route. Automated testing is performed under the control of a central minicomputer at access points provided in both the multiplex and the protection switching equipment. Both in-service and out-of-service testing capabilities are provided with special stress-testing features implemented at radio repeater locations for use in trouble isolation and diagnosis. The computer-controlled measurement system is designated "Transmission Surveillance System—Radio" and is described in detail in this issue of the *Journal.*

IV. SINGLE-SIDEBAND HISTORY

In the late 1960s, detailed system studies were performed by A. J. Giger to define linearity requirements for radio repeaters (TR units) for long-haul single-sideband transmission. Several methods were recommended for achieving the necessary transmitter performance. Subsequently, experiments were performed in the early 1970s with feedforward microwave techniques to improve the linearity of traveling-wave-tube amplifiers, the most difficult challenge in achieving the radio transmitter intermodulation noise requirement. The success of these experiments led to addressing the problems of overall repeater linearity, equalization for multipath fading effects, and the effects of interference when used in the existing FM radio network.

Studies were performed to obtain a better understanding of traveling-wave-tube distortion, and this resulted in the use of predistortion to meet transmitter linearity requirements. Additional field data were gathered on multipath fading, and these data were a key ingredient in achieving a simplified dynamic equalization design. Supporting studies of a subjective nature also were conducted to provide more detailed

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

definitions of tone and crosstalk annoyance objectives when Single-Sideband Amplitude Modulation (SSBAM) was used in the radio network; the frequency control plan and the intermediate frequency (IF) filter design were chosen to meet these objectives. The adequacy of the methods for linearization, equalization, and interference control were established by a field test performed in the 4-GHz network in 1974 and early 1975.

Detailed system design of AR6A for the 6-GHz network started in 1975, and in 1979 a complete system was installed on a 6-hop radio route from Hillsboro, Missouri, to Windsor, Missouri. All subsystems of AR6A were tested and evaluated on this 177-mile route, and system verification testing was completed on June 14, 1980. This route then was extended by three radio hops (71 miles) to La Cygne, Kansas, and became the first single-sideband amplitude modulation route to be placed in commercial service on January 12, 1981.

V. ACKNOWLEDGMENTS

The feasibility demonstration of SSBAM transmission on the long-haul 4-GHz radio network in 1974–1975 was the culmination of the efforts of many people in different organizations throughout Bell Laboratories. The realization of AR6A for the 6-GHz network was the result of the dedicated efforts of many organizations at Bell Laboratories working with Western Electric and in close concert with AT&T Communications. Technology foundations were provided by the TD, TH, and L5 Systems and these were augmented by advances in filter, network, and device designs, as well as the effective use of the many computational and measurement facilities available at Bell Laboratories. The authors of this volume express their appreciation to all those who have contributed to the ultimate attainment of SSBAM transmission on the long-haul radio network.

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The AR6A Single-Sideband Microwave Radio System:

System Design and Performance

by J. GAMMIE,* J. P. MOFFATT,* R. H. MOSELEY,* and W. A. ROBINSON*

(Manuscript received February 14, 1983)

This paper describes the overall architecture of the first long-distance microwave telephone transmission system to use Single-Sideband Amplitude Modulation in the microwave transmission path. Compared with commonly used frequency modulation systems, the AR6A Single-Sideband Microwave Radio System more than doubles the number of telephone circuits that can be carried per unit of microwave bandwidth. Included in the description are the system design model, performance objectives, and the allocation of performance impairments. Also discussed are the computer-controlled automated maintenance features and the results of laboratory and field tests.

I. GENERAL DESCRIPTION

1.1 Introduction

The AR6A† System is the first long-haul microwave transmission facility to employ Single-Sideband Amplitude Modulation (SSBAM)‡ in the microwave transmission path. To fully exploit this alternative to the widely used frequency-modulation (FM) system, it has been necessary to develop for SSBAM new subsystems to accomplish the

* Bell Laboratories.

† 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.*

functions commonly associated with an integrated microwave transmission facility. In this sense, the use of SSBAM affects not only the design of the transmitter-receiver (TR) equipment but all ancillary equipment as well.

Broadly speaking, the main functional elements of the system are fairly conventional, but a number of SSBAM's basic requirements have had a major impact on the AR6A System design. The first and most obvious of these is the high level of linearity needed in many of the transmission elements to control intermodulation distortion. Although the most taxing task was making the transmitting traveling-wave-tube amplifier meet overall system linearity objectives, the intermodulation performance of many other elements required careful scrutiny and improvement. Frequently, the more stringent linearity requirements have been associated with higher quiescent current for transistors and, in general, with increased power requirements relative to comparable elements in an FM system. Nevertheless, when power requirements are normalized to a per-circuit or a per-mastergroup basis, power utilization efficiencies are very competitive.

In addition, the direct dependence of signal amplitude response on the response of the transmission medium is particularly stressing during multipath fading. This requires adding a capability for dynamic amplitude equalization. Design of the necessary adaptive equalizers required an extensive experimental effort aimed at quantitatively describing the statistics for and understanding the effects of multipath fading on channel amplitude shape. The dynamic dispersive channel resulting from multipath fading required not only adaptive equalizers but also the use of space diversity.

Unlike FM, in which demodulation is performed relative to a transmitted FM carrier, accumulated frequency shifts due to multiple modulation and remodulation at repeaters must be removed. This adds to the complexity of the terminal multiplex equipment that performs the frequency-correction function.

Finally, successful transmission of SSBAM requires new maintenance and equalization considerations. In FM systems, minor amounts of transmission slope and absolute amplitude changes do not greatly affect the quality of the recovered baseband signal. The greater sensitivity of SSBAM to these effects, particularly their potentially cumulative nature, requires much closer control of equalization and absolute amplitude levels. These considerations, combined with the trend towards centralized operational support systems, resulted in the integrated development of a minicomputer-based Transmission Surveillance System—Radio (TSS-R) with associated trouble isolation and stress test features.

During development, effort was initially focused on applications employing frequency-diversity protection, the most commonly used system arrangement. This work was later supplemented by the development of hot-standby and space-diversity arrangements for applications where lower cross-sectional growth rates preclude the use of frequency-diversity protection.

Because of the many new system features, AR6A underwent an early but limited field trial between Ashburnham and Wendell, Massachusetts, in the period October 1977 to July 1978. This two-hop installation provided valuable initial field experience prior to a more comprehensive field evaluation between Hillsboro and Windsor, Missouri. The Missouri evaluation involving a six-hop radio section with frequency- and space-diversity protection switching encompassed the period from April 1979 to June 1980. The evaluation of hot-standby arrangements was performed later on a two-hop section between Colorado Springs and Cedarwood, Colorado. The shipment of standard AR6A equipment began in June 1980; first commercial service was placed on the AR6A route between Hillsboro, Missouri, and La Cygne, Kansas, in January 1981.

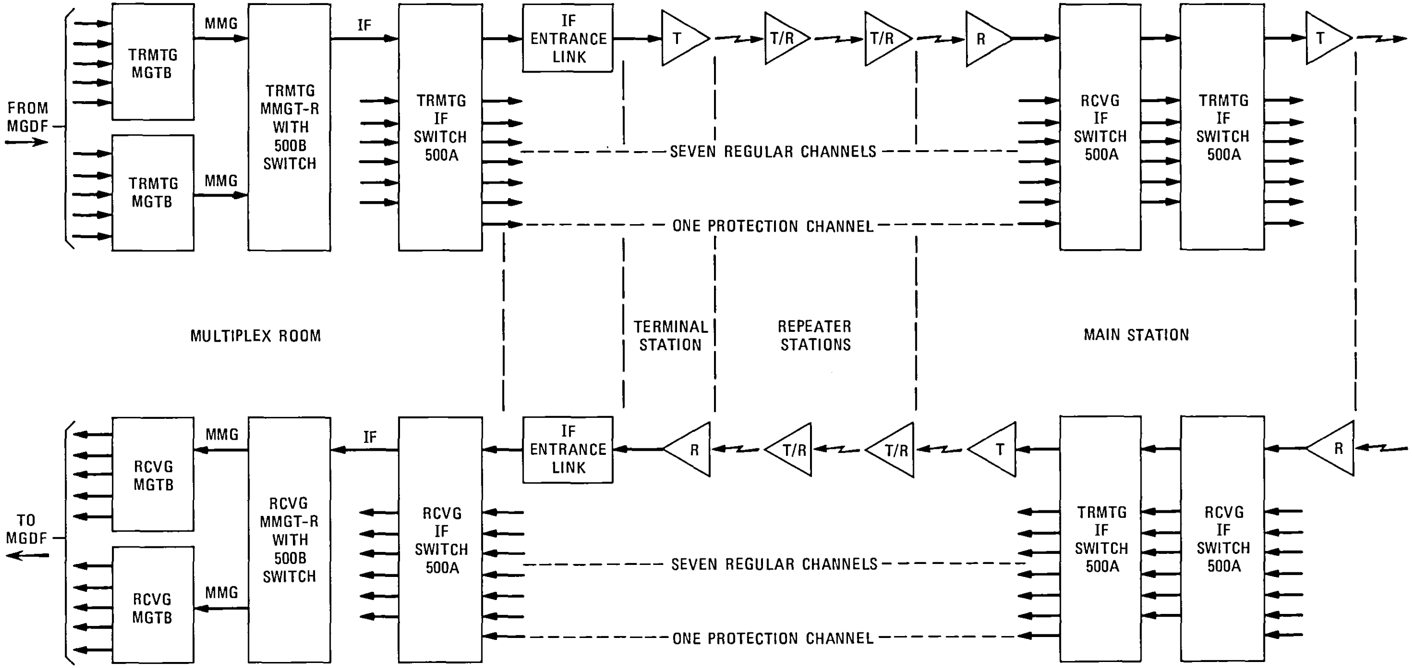
The following paragraphs are devoted to an overview of the AR6A System using the block diagram of Fig. 1. The figure and discussion relate to a system configuration using frequency-diversity protection switching which, besides being the most common application, will provide a satisfactory medium for describing the main features of the system.

1.2 Multiplex

1.2.1 Mastergroup translators

The AR6A System interfaces with the standard Bell System analog multiplex hierarchy at the basic U600 mastergroup level. This is also the highest level in the multiplex hierarchy at which AR6A can be interconnected to other broadband transmission facilities. The U600 mastergroup contains 600 4-kHz circuits in the frequency band 564 to 3084 kHz. Figure 1 illustrates how the basic mastergroup is delivered from the Mastergroup Distributing Frame (MGDF) to the first frequency translation step provided by the Mastergroup Translator, Type B (MGTB). Five MGTBs translate and frequency multiplex five separate mastergroup signals into adjacent positions in a multimastergroup band extending from 8.628 to 21.900 MHz. With a fixed intermastergroup spacing of 168 kHz, the tight packing provided by the MGTB is an important factor in being able to place two multimastergroups (ten mastergroups) in each 30-MHz broadband radio channel.

In addition to modulation and demodulation, the MGTB modems supply the 2840-kHz mastergroup pilot which is used in the receiving



MGDF – MASTERGROUP DISTRIBUTING FRAME
MGTB – MASTERGROUP TRANSLATOR, TYPE B
MMG – MULTIMASTERGROUP
MMGT-R – MULTIMASTERGROUP TRANSLATOR-R
RCVG – RECEIVING
TRMTG – TRANSMITTING

Fig. 1—Block diagram of AR6A System.

portion of the modem for gain regulation purposes. Other functions associated with the MGTB combining circuits are (1) means for inserting a 13.920-MHz multimastergroup continuity pilot for sectionalizing troubles, (2) a comb filter with narrowband band-stop sections which remove signals and noise from intermastergroup slots prior to the insertion of line pilots, and (3) test access points for TSS-R and other automated measuring facilities.

1.2.2 Multimastergroup translators

The Multimastergroup Translator—Radio (MMGT-R) performs the next modulation step by translating two multimastergroups into the intermediate frequency (IF) band centered at 74.13 MHz. The disposition of the multimastergroups and the previously described translation steps from basic mastergroup are illustrated in Fig. 2.

In the MMGT-R, the two multimastergroups are translated to IF by separate modulators with independent, unsynchronized, crystal-controlled local oscillators. On the receiving side, the demodulating carriers are supplied by Voltage-Controlled Oscillators (VCOs). Each VCO tracks the accumulated line frequency errors with a Phase-Locked Loop (PLL) to deliver a frequency-corrected multimastergroup signal at the receiving end of the terminal section. Essential to the frequency-correcting process is the addition of an intermastergroup recovery pilot to each multimastergroup signal before it is translated to IF. This pilot at 16.608 MHz is frequency locked to a 512-kHz synchronizing signal derived from the Bell System Reference Frequency (BSRF). In view of the recovery pilot's importance, the synchronizing source supplied to the MMGT-R is duplicated and automatically protected. At the output of the receiving MMGT-R, the demodulated recovery pilot is compared with a locally generated reference frequency to provide the PLL error signal for frequency correction. An important additional function of the receiving PLL is to reduce accumulated phase jitter on the radio line.

In addition to recovery pilots, the transmitting portion of the MMGT-R supplies three radio-line pilots from free-running crystal oscillators. When translated to IF these pilots are located near the center and band edges of the radio channel, as illustrated in Fig. 2. At radio receivers, the pilots provide a reference signal for Automatic Gain Control (AGC) and, in addition, provide information on channel amplitude shape for dynamic amplitude equalization. The radio pilots, recovery pilots, and multimastergroup pilots are each transmitted at a level of −10 dBm0.

1.2.3 Office master frequency supply

Carrier supplies for the Bell System analog multiplex hierarchy are

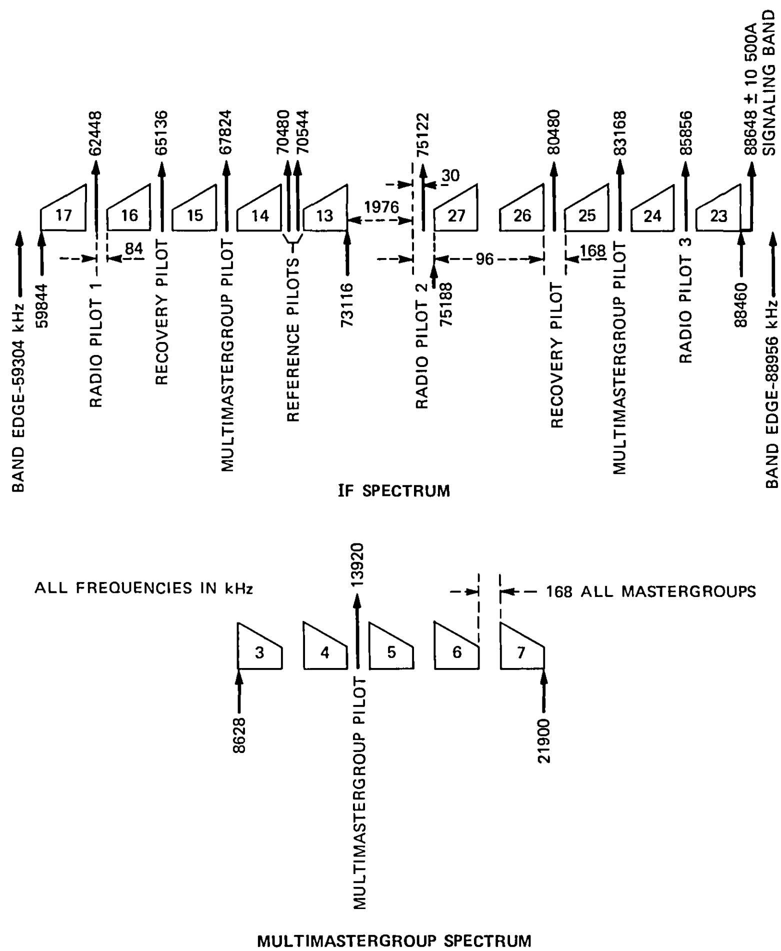


Fig. 2—Multimastergroup and IF spectrum.

locked to a 64- or 512-kHz synchronizing source derived from an office Primary Frequency Supply (PFS). The PFS is in turn synchronized to a signal with a frequency accuracy that is traceable within very narrow limits to that of the BSRF source at Hillsboro, Missouri. If the PFS temporarily loses its input synchronizing signal, the output frequency will suddenly change to the unit's natural or rest frequency. This frequency transient, although tolerable at lower levels in the multiplex hierarchy, was considered excessive in a synchronizing source for the AR6A recovery pilots. For this reason, a new Office Master Frequency Supply (OMFS) was developed which provides 64- and 512-kHz synchronizing signals that are quasi-frequency locked to the 2.048-MHz BSRF. With quasi-frequency lock, instead of being synchronized to the incoming reference signal, the highly stable OMFS oscillators are free-running. Their frequency is compared with the reference frequency at intervals and smoothly updated if the error exceeds the prescribed threshold. If the reference frequency is temporarily lost, updating is inhibited and the oscillators continue to run at their most recently updated frequency. In this way the OMFS shields dependent multiplex equipment from transient effects on the synchronizing network.

1.2.4 Reference frequency transmission

The 2.048-MHz Bell System Reference Frequency must be available at all AR6A terminal stations for the purposes described in Section 1.2.3. The reference signal can be made available over a variety of transmission facilities, but there are instances where the only available transmission medium will be the AR6A System. To provide this capability, an optional reference frequency sending unit is available that inserts two pilots between MG13 and MG14 (Fig. 2) at a level of −15 dBm0. The pilot frequencies differ by 64 kHz with an accuracy determined by the sending end BSRF. At the receiving end of the terminal section, a reference frequency receiving unit recreates a BSRF at 2.048 MHz from the accurately defined 64-kHz difference frequency. In this way, reference frequency transmission accuracy is unaffected by inaccuracies in the multiple frequency translations experienced on the radio line. Sixty-four-kHz rather than a wider frequency separation was chosen to minimize transmission dispersion effects and to optimize pilot positions from the standpoint of noise and freedom from interaction with other system pilots. The sending and receiving frequency units are physically associated with the MMGT-R equipment.

1.3 The radio line

1.3.1 Entrance links

Figure 1 shows that the IF signal from the MMGT-R is fed to the radio transmitting equipment via the 500A Protection Switching System, which will be discussed in Section 1.4. Since the radio transmitters may be at some physical distance from the preceding multiplex equipment, provision must be made for handling trunk losses and associated transmission shapes. In some instances, the trunks may be so long that they require amplification, necessitating automatic protection of the active circuits. In the AR6A System, the 500A protection switching equipment is located near the MGTB and MMGT-R multiplex equipment. In this way, the trunks or entrance links become extensions of the radio line and may, therefore, share the 500A protection facilities.

For distances up to 600 feet, the entrance link trunks are passive and are equalized to within 25 feet. Greater distances require amplification by amplifiers located in the TR bays for both the transmitting and receiving trunks. This provides for entrance link distances of up to 1200 feet using 728-type office coaxial cable. Entrance link distances up to 3600 feet can be accommodated with 0.375-inch coaxial cable.

1.3.2 Transmitter-receiver equipment

The heterodyne-type transmitter-receiver (TR) bays are designed for overbuilding on existing 4-GHz radio routes with an average repeater spacing of 25.6 miles. The nominal repeater gain is 61.7 dB with up-fade and down-fade ranges of 15 and 46 dB, respectively. The full-load transmitter output power is +24.6 dBm, reflecting an average power per 4-kHz message circuit of −19.6 dBm0.

The TR bay equipment is designed to permit field installation by technicians into existing bay frameworks. This modular installation allows the bay frameworks to be installed and the station wiring to be done for all radio channels at one time. The TR modules are installed as required by the route traffic.

The transmitter uses IF predistortion to compensate for the nonlinearities of the up converter and the output traveling-wave-tube amplifier. Automatic gain control (AGC) at each receiver is based on maintaining the voltage average of the three pilots constant at the receiver output. In addition to AGC, the last receiver in each switch section is equipped with a dynamic amplitude equalizer.

A microwave preamplifier with a noise figure of approximately 3 dB is provided in the common receiving waveguide run on each polarization. If the hop length is less than 63 percent of nominal, the preamplifier is omitted.

To meet long-haul reliability objectives within the FCC-prescribed limit of only one frequency-diversity protection channel to serve up to seven service channels, it is estimated that approximately half the radio hops will require space diversity. The TR bay is, therefore, equipped with an optional space-diversity switch that is actuated by a 36-dB fade in any one of the three radio-line pilots.

In the event of a loss of received signal due to equipment trouble or other reasons, a pilot resupply signal is automatically inserted at the receiver output. In addition to supplying the three radio-line pilots, the resupply source provides a fourth pilot that is located in the center noise-monitoring slot of the 500A Protection Switching System. The channel with resupply is thereby marked noisy and unavailable for service. The pilot resupply signal at each station is provided from a common source and distribution network located in a radio support bay.

Networks in the IF shelf provide repeater bandpass shaping and basic equalization for the nominal repeater transmission shape. Also, positions are provided for limited amounts of mop-up equalization for nonsystematic amplitude response variations.

Carrier supplies for up and down frequency conversion are provided by the TR bay microwave generator and its associated 252-MHz shift oscillator. To maintain pilot frequencies within the passband of the narrowband pilot pick-off filters at each repeater, the carrier frequencies must be accurate to better than one part in 10^7. This is achieved by locking the microwave generators and shift oscillators to a common submultiple frequency, 308.87354 kHz. This synchronizing frequency is distributed to each TR bay from a common synchronizing source located in the radio support bay. If the synchronizing signal fails, both the microwave generator and shift oscillator have memory circuits that maintain their frequencies at the most recently updated values. Subsequent drifting is at a rate that allows a one- to two-day window for repairing the failed synchronizing source.

1.3.3 Support bay

The support bay contains the Microwave Carrier Synchronization Supply (MCSS) and the pilot resupply source, along with associated distributing circuits. In addition, it provides power distribution, fusing, and alarm arrangements for up to eight microwave preamplifiers. The support bay is designed to supply up to 16 TR bays corresponding to one fully loaded route passing through a repeater station. In hot-standby applications, the radio support bay also contains the hot-standby control unit.

1.3.4 Antenna system

Since a common application of AR6A will involve overbuilding on existing 4-GHz radio routes, the system is designed around the currently used horn antenna system with associated system combining and separation networks. The space-diversity path should essentially duplicate the main path for a number of reasons, including the anticipated later application of space-diversity combining. With nominal tower heights and hop lengths, the combined section loss, waveguide losses, network losses, and antenna gains give a net overall nominal section loss of 61.7 dB.

1.4 Protection switching

1.4.1 500B Switching System

Since each MMGT-R carries 6000 message circuits, automatic transfer of the message load to a hot-standby protection unit is necessary for reliability and maintenance. The function is handled by the 500B System. One microprocessor-based controller oversees a 1 ×

14 protection arrangement for transmitting MMGT-Rs and a corresponding 1 × 14 protection arrangement for receiving MMGT-Rs. Since the protection function is local, multimastergroup translators from more than one route may be associated within the same 1 × 14 switching system.

In its automatic mode, a switch is initiated by a number of conditions including loss of the multimastergroup continuity pilot and loss of synchronization.

1.4.2 500A Protection Switching System

The 500A System provides frequency-diversity protection switching on the radio line. Switching sections may be up to ten hops in length and the microprocessor-based control functions are located at the receiving end. The protection switching systems in the two directions of transmission are electrically independent.

Each protection channel can serve up to seven working channels. A frequency shift-keyed (FSK) tone transmitted over the oppositely directed radio channels signals between the receiving and head end of the switch section. The FSK tone is inserted at IF near the upper channel edge and is blocked at the receiving end of each switch section.

Switch initiators monitor each of the three radio-line pilots as well as the noise in three narrowband slots adjacent to the pilots. This multiple monitoring causes more switching activity than the single-point monitoring on FM systems and is partly responsible for the approximately 50-percent space-diversity requirement for AR6A. In addition to protecting against frequency-selective fading, particularly on hops without space diversity, 500A protects against equipment failure and provides out-of-service access to the radio channels for maintenance purposes.

Maintenance access to a radio channel via a maintenance switch command to the 500A System is an essential element in the implementation of the centralized Transmission Surveillance System—Radio (TSS-R). Besides providing switched access to a channel at the transmit and receive ends of the switch section, the 500A control logic is responsible for ensuring that accessing the channel is subordinate to maintaining service continuity. Even after out-of-service access to a channel is permitted, a preemption feature in the 500A logic will restore the channel to service automatically if this is necessary to prevent a service outage. When a channel is opened at the head end of the switch section to allow maintenance access, a fresh set of locally generated radio pilots must be inserted to provide proper control at subsequent radio receivers. These pilots are provided by the 500A System in the same way as resupply pilots are provided at repeaters.

1.5 Hot standby

On routes where the cross-sectional growth rate is low, current FCC rules do not permit the use of frequency-diversity protection channels. In these circumstances, protection against outage due to multipath fading is generally provided by using space diversity. To protect against radio equipment outage, space diversity must be supplemented with hot-standby radio equipment that is automatically switched into service when the need arises. This hot-standby configuration is available on AR6A although the number of applications is expected to be small compared to the frequency-diversity configuration.

Since radio repeaters are frequency-dependent, each working TR bay generally has a separate hot-standby bay. In this sense, hot-standby configurations are equipment-intensive. To ease this situation in AR6A, the east-west and west-east repeaters on a given radio channel are identical, including the frequency-dependent components. By providing appropriate switching and control arrangements, a common TR bay can be used to protect equipment for both directions of transmission. Main stations do not lend themselves to this reduced configuration and are, therefore, arranged with conventional one-for-one hot standby.

Hot-standby manual and automatic control functions at repeaters are provided by a microprocessor-based control unit. The same unit with a subset of features is used at main stations to control hot-standby switching of transmitters. Main station receivers are protected by the 501A Protection Switching System, which also has a double feed to the regular and hot-standby transmitters. In addition to providing noise and pilot-level monitoring with 500A-type initiators, the 501A System also enables the TSS-R System to monitor in service.

On a hot-standby route, each repeater has its own protection against equipment outage and multipath fading. There is, therefore, no natural equivalent to the multihop switch section of the frequency-diversity configuration. However, since hot-standby routes will often be converted to frequency diversity when growth permits, main station TR bay and waveguide configurations are recommended at those stations which at a later time may become switching or main stations. Since the main station configuration implies the use of the 501A switch with surveillance access, this sectionalizing of hot-standby routes into surveillance sections subdivides the route naturally for purposes of dynamic equalization, transmission monitoring, and trouble isolation. The interface transmission levels on the 501A System are identical to those on the 500A switch to further simplify the transition from hot standby to frequency diversity.

1.6 Maintenance

Compared with FM systems, the somewhat less rugged modulation technique and the higher capacity of the AR6A System have emphasized the need for careful maintenance planning in the overall system design. Supplementing these considerations is the ongoing trend towards a demand maintenance scenario, improved trouble isolation procedures, and centralized transmission surveillance capabilities.

On AR6A, status reporting and command functions are handled conventionally utilizing the C1 or E2 systems in conjunction with the Surveillance and Control of Transmission Systems (SCOTS) or Telecommunication Alarm Surveillance and Control (TASC) central processors. However, in the 500A-to-E2 interface, which can involve a sizeable number of wire pairs, optional arrangements have been developed to provide a serial interface with only two wired pairs. This feature capitalizes on the 500A microprocessor that can communicate serially to microprocessor-based alarm remotes such as E2A.

For trouble-isolation purposes, a significant new feature is the ability to break the transmission path and insert resupply pilots at any selected station by remote command. This is supplemented by the capability to remotely modify the resupply signal to perform a variety of stress tests. With this feature, in conjunction with a measuring capability at the receiving end of a switch section, a variety of transmission troubles can be localized by making out-of-service measurements on successively shortened portions of the switch section. Of course, these capabilities must be circumscribed with protections against service interruption and loss of the 500A preemption feature. These matters are discussed more fully in Section 4.9.

Another important new maintenance feature is the Transmission Surveillance System—Radio (TSS-R). This system uses programmable measuring, signal source, and access equipment at radio main stations that are under the control of a local microprocessor. This microprocessor is, in turn, accessible and controlled via the Direct Distance Dialing (DDD) network from a central minicomputer. Data link dial-up and communication protocol, measuring, evaluation, and reporting routines residing in the central software provide for a variety of switch section transmission measurements that can be made either on an in-service or out-of-service basis. Furthermore, by allowing the TSS-R central minicomputer to talk to a SCOTS or TASC central, the trouble-isolation and stress test features described in the preceding paragraph can be integrated into automated programmable procedures. It should again be noted that access to a radio channel for in-service or out-of-service measurements is subject to approval by the 500A System control logic. It is anticipated that five TSS-R centrals will be able to cover all AR6A routes in the continental United States.

With respect to equipment maintenance, test equipment is, in general, conventional. However, transmitter linearization and precise frequency adjustment of the MCSS require some special test equipment, which is described in a companion paper.[1]

II. PERFORMANCE OBJECTIVES

2.1 Noise

The 4000-mile, voice-circuit noise objective is 40 dBrnc0 under standard (free-space) propagation conditions. Attaining this level of performance requires state-of-the-art designs for many devices in the system and the recognition that the current per-circuit average power in the telephone network is lower than previous designs had assumed. The per-circuit average power used in the system design is −19.6 dBm0. This level has been found to be representative of both radio trunks[2] and the telephone network as a whole.[3] With the current increase in the use of common channel interoffice signaling, it is expected that this average power will decrease slightly.

2.2 Amplitude equalization

2.2.1 Static equalization

The static amplitude equalization objective requires the amplitude shape developed across a mastergroup to be not greater than ±3 dB in 4000 miles. This limit is established to be within the available correction range of regulators in the receiving multiplex equipment. Typical regulators have a ±6 dB range and ±3 dB is chosen as the allocation for the radio system. A supplementary objective of ±0.5 dB is established for the total misalignment within a radio channel in any given switch section. This limit prevents excessive errors in voiceband data systems during frequency-diversity switching between the regular and protection radio channels.

2.2.2 Dynamic equalization

As a result of the time-variant nature of propagation over typical microwave radio paths, an objective for time-variant signal amplitude is established. The objective is that for a 4000-mile circuit, the amplitude error should be less than ±2 dB for 99.9 percent of the time. The ±2 dB limit is chosen since variations exceeding this level can cause errors in some voiceband data systems. Ordinary telephone service is more tolerant of this impairment.

2.3 Tone interference

The tone interference objective uses the ratio of spurious tone power to background (masking) noise power in a voice circuit as the measure of impairment. This measure is directly related to the annoying effect perceived by the customer. It also has the desirable feature of being straightforward to administer in a large system, since each subsystem produces its own background noise and the tone requirements for the individual subsystems follow directly. The objective is stated on a percent of time basis as follows. The tone-to-noise power ratio should be less than: (1) −10 dB for 99 percent of the time, (2) 0 dB for 99.9 percent of the time, and (3) 2 dB 100 percent of the time.

In applying the tone objective to the radio line, it is found that on long routes the radio-line noise is dominated by the noise of a faded hop for more than 1 percent of the time. For tone mechanisms that increase with fading, the receiver thermal noise is then used as the background noise for the per-repeater tone objective. For tones that do not increase in level, the total radio repeater noise is used.

2.4 Crosstalk

Separate objectives have been established for intelligible and unintelligible crosstalk. The objective for the former is that the crosstalk index should be less than 0.5 percent. This corresponds approximately to a probability of 0.005 of hearing one or more syllables of intelligible crosstalk during an average duration call. The unintelligible crosstalk or babble objective is stated in terms of a babble-to-noise power ratio that must be met during certain percentages of the time. The babble-to-noise ratio must be less than −10 dB 99 percent of the time and less than 7 dB 100 percent of the time.

2.5 Phase jitter

The objective for phase jitter (unwanted phase modulation) in 4000-mile circuits is 8 degrees peak to peak. The objective applies to jitter with either discrete or continuous frequency components. The jitter is measured by a standard jitter indicator[4] that is sensitive to spectral components in the range of 20 through 300 Hz.

2.6 Reliability

The reliability objective for a 4000-mile circuit is that it should be available for two-way service with noise less than 55 dBrnc0 for 99.98 percent of the time. This objective is apportioned among three sources of outage as follows: (1) equipment outage, 0.005 percent; (2) flat fading, 0.005 percent; and (3) selective fading, 0.010 percent. To meet the equipment reliability objective, the system design requires either frequency-diversity or hot-standby protection switching. The application of space-diversity antennas to meet the selective fading objective is discussed in Section 4.7.

III. SYSTEM MODEL AND PERFORMANCE ALLOCATIONS

3.1 The 4000-mile system model

As a basis for calculating the expected performance of the radio system, a 4000-mile system model is used. The model is derived from the characteristics of the existing Bell System microwave radio network. The network has a mean hop length of approximately 25.6 miles, resulting in 156 hops in 4000 miles. The network root-mean-square (rms) hop length is 27.1 miles and this latter value is used for thermal noise and received signal-level calculations. The section loss corresponding to the rms hop is assumed to be 61.7 dB with typical antenna systems. This value includes all losses and gains from the output monitor-shutter in the transmitter to the input of either the receiving bay lineup or the microwave preamplifier if it is installed. The 4000-mile route is assumed to have 52 switch sections averaging three hops in length. Multiplex and terminal equipment is assumed to be located at nine intermediate points on the route. A 4000-mile circuit is thus assumed to pass through ten sets of MMGT-R and MGTB equipment. In addition, the circuit is assumed to pass through a total of nine sets of supergroup multiplex, six sets of group multiplex, and three sets of channel banks. The mastergroups within the spectrum of a radio channel are assumed to be interchanged or frogged every 400 miles for six of the terminal sections and 800 miles for two of the terminal sections.

3.2 Performance allocations

3.2.1 Noise

Figure 3 shows the noise allocation for the system. The terminal equipment allocation results from assuming two-thirds of the multiplex equipment below mastergroup level is LMX-2 and one-third is LMX-3. This mix is representative of the network during the initial introduction of the system. Thermal and intermodulation noise are the primary contributors to the terminal equipment and protection switching allocations.

Four major interference mechanisms contribute to the RF interference allocation. Cochannel same-route exposures occur at the nine intermediate multiplex stations in the system model. At other stations, the same-route mechanism produces talker echo as a result of the system frequency stability and Radio Frequency (RF) channel plan. At stations with intersecting routes, cochannel junction interference occurs. The allocation assumes a mixture of AR6A and FM intersecting routes that will exist during the initial introduction of AR6A into the radio network. The third mechanism is cochannel interference from other terrestrial radio systems. All stations in the system model are subject to this interference. The final mechanism is cochannel interference from geostationary satellite earth stations. An allocation of 30 dBrnc0 has been set for this source.

The major contributors to the radio repeater allocation are thermal

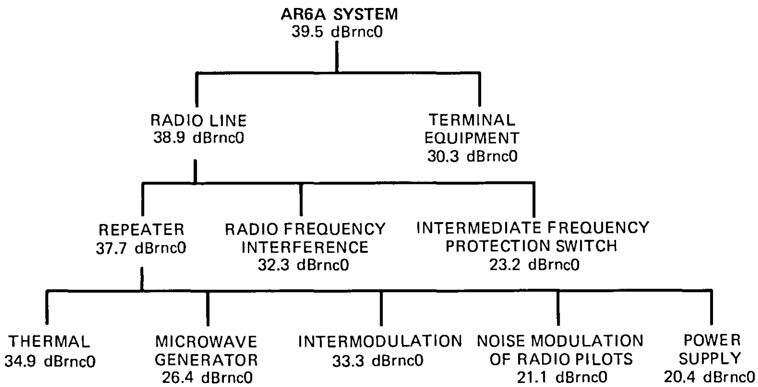


Fig. 3—AR6A noise allocation.

and intermodulation noise. The relative levels of these contributors result from an optimization of the overall system noise with respect to transmitter output power. During the development process, the allocations for these sources were lowered as a result of the use of the microwave preamplifier and the lower per-circuit average power of −19.6 dBm0.

The microwave generator allocation results from phase noise sidebands produced by the generator. In the modulation processes that use the generator output, these sidebands are convolved with the signal and produce an unwanted additive noise. In the system design, this noise is minimized at repeater stations by using one microwave generator to down-convert and up-convert each channel. The noise introduced in the two conversions nearly cancels. The predominant contributors to the allocation are the separate microwave generators used at main stations.

Since the system gain is controlled by the detected level of the three radio pilots, any spurious energy entering the passbands of the detection circuitry causes unwanted amplitude modulation of the signal. The modulation results in additive noise. The allocation for noise modulation of the radio pilots results from accumulation within a multiplex section of radio-line thermal and other noise near the radio pilots. At multiplex locations, the noise from previous paths, within the detection circuitry bandwidth, is removed and new pilots are inserted.

The power supply allocation results from a similar process. Ripple or other nonuniformity of the power supply voltages results in unwanted modulation of the signal. That portion of the modulation energy that is outside of the radio pilot detection circuitry bandwidth remains on the signal and produces noise. The predominant contributor to this allocation is the traveling-wave-tube (TWT) power supply.

3.2.2 *Amplitude equalization*

The static equalization objective for mastergroup shape is allocated among: (1) the residual shape of IF and MMG trunks, (2) the residual shape of an equalized radio bay, (3) echoes in the antenna and waveguide systems, (4) signal leakage through the space-diversity switch, and (5) tertiary interference. The predominant shape component is slope. The slopes introduced by all of the above sources are random and contribute to an estimated rms slope of 1.21 dB in 4000 miles. The slope distribution is approximately normal and the probability of exceeding a slope of 3 dB is then 1.3 percent. This probability is sufficiently small to satisfy the intent of the objective.

3.2.3 *Crosstalk*

The main sources of crosstalk in AR6A are interference exposures at junction stations and multiplex locations. The type of crosstalk produced by these exposures depends on the relative frequency stability of the interfering and desired signals. The calculated probabilities for the interference exposures producing intelligible crosstalk and babble are 0.45 and 0.55, respectively. Calculations using interference coupling levels typical of the radio network result in a crosstalk index of 0.12 percent for 4000-mile routes. The calculated probability of exceeding a −10 dB babble-to-noise ratio is 0.32 percent for 4000-mile routes. The remainder of the crosstalk and babble objectives are allocated to mechanisms producing discrete-frequency incidental modulation in the radio line.

IV. DESIGN ASPECTS

4.1 *Frequency plan*

4.1.1 *Channel assignments*

The AR6A System uses the standard 6-GHz frequency plan, which is shown in Fig. 4. This plan permits eight channels in each direction of transmission, which are designated the same as in the TH System: 11 through 18 in the lower portion of the frequency band and 21 through 28 in the upper portion of the frequency band. This plan places the receivers in one-half of the band and the transmitters in the other half. A station using the frequency plan shown in Fig. 4 is called a "low-high" station, because the receive channels are in the low-frequency portion of the band and the transmit channels are in the high-frequency portion of the band. Adjacent stations will be "high-low" stations. The microwave carrier frequencies, which are used in the receiver and transmitter modulators, are also shown in the figure, along with the channels served.

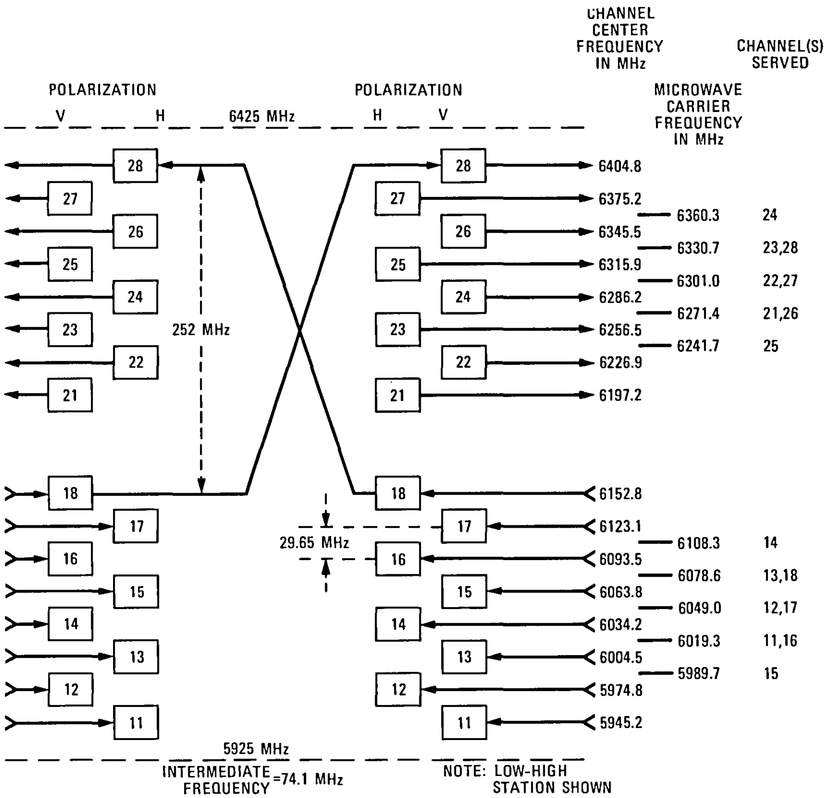


Fig. 4—AR6A frequency plan.

4.1.2 Choice of IF

An IF band centered at a frequency that is an odd multiple of half the channel separation is the best choice for minimizing the effects of microwave carrier leakages. With this choice of IF center frequency, these leakages appear at the edge between two channels and do not need as much suppression as they would if they fell inside a channel. On this basis the best IF center frequency for the 6-GHz frequency plan is 74.12965 MHz (usually referred to as 74.1 MHz). In addition to the advantage of controlling the effect of the microwave carrier leakages, a 74.1-MHz center frequency permits a reduction from 16 to 10 of the microwave carrier and filter codes needed, since some microwave carrier frequencies can be used for two channels by using upper and lower sideband modulators. These design factors favor the choice of 74.1 MHz. Consideration was given to using a center frequency of 70.0 MHz, which is used in the TH-3 and TD Radio Systems. This choice would permit the direct IF interconnection of FM systems onto an AR6A radio channel. The ability to handle both AM and FM signals would have excessively complicated the AR6A design; therefore, an IF center frequency of 74.1 MHz was used.

4.1.3 IF band utilization

At both IF and microwave frequencies, the SSBAM signal with suppressed carrier occupies no additional bandwidth over the original baseband signal. Thus, mastergroups can be translated into the channel bandwidth with a spacing determined by the filtering requirements needed for separation when shifted back to baseband. Carrier systems previously used a frequency spacing between mastergroups of approximately 4 percent of the frequency, where the mastergroup was to be located based on the desire to separate mastergroups (block and branch) at the multimastergroup level. The use of mastergroup translators makes it more convenient and economical to make these rearrangements at the basic mastergroup level, which permits the use of a fixed frequency separation at the multimastergroup level. Improvements in filter performance have made it possible to form the MMG spectrum with only 168-kHz spacing between mastergroups. The number of mastergroups that can be placed in the radio channel bandwidth is constrained by cochannel interference from FM systems. Clearly, the center of the channel, which may contain interfering FM carriers, cannot be used, and further allowance must be made for the close-in sidebands of the FM signal since these are strong enough to cause serious interference into AR6A voice circuits. The situation for TH-3 is shown in Fig. 5. Consequently, a center gap of 2.072 MHz has been provided which, with frogging of the mastergroups, keeps interference at an acceptable level, and achieves a loading of ten mastergroups in the 29.65-MHz channel bandwidth. The addition of pilot tones in the gaps between mastergroups for regulation and control purposes completes the IF spectrum as shown in Fig. 2.

4.2 Frequency stabilization

4.2.1 Need for stabilization

An important aspect of the system design is the required frequency stability along a radio route. The radio pilots, which are transmitted as part of the IF spectrum, must be centered in their respective pickoff filters in each receiver. For best operation, the pilots should be within ± 2 kHz of their nominal frequencies. The necessary frequency accuracy is achieved by locking the microwave generators and shift oscillators at each station to a high-stability synchronization oscillator located in each station. This will stabilize both the microwave and the IF frequencies of the system.

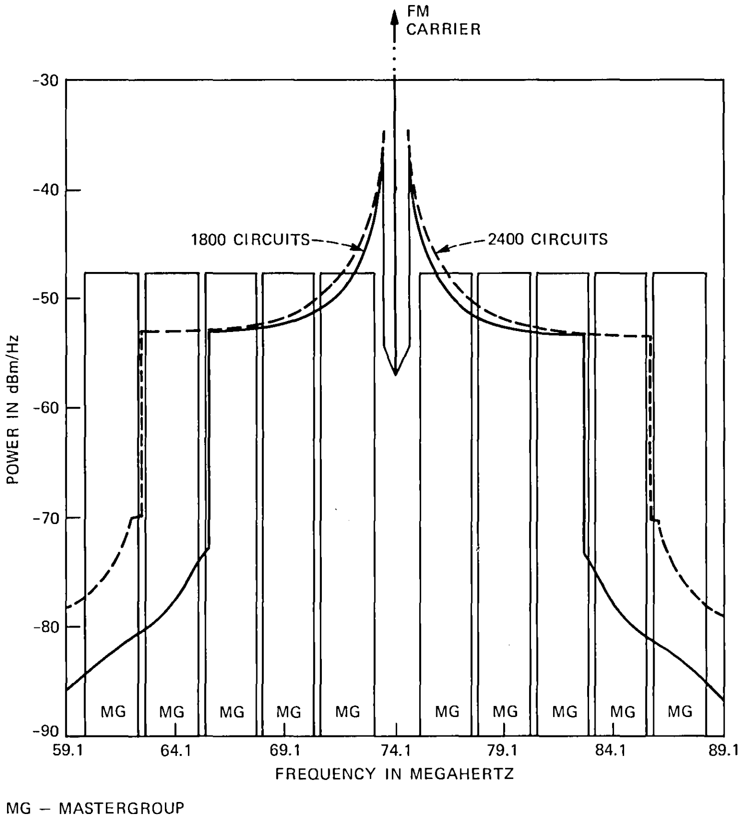


Fig. 5—AR6A and TH spectra.

4.2.2 Method of stabilization

As a result of choosing the IF center frequency of 74.12965 MHz, all the microwave generator frequencies and the shift oscillator frequency are multiples of 14.825930 MHz. Because the best crystal frequency stability is obtained around 5 MHz, the AR6A synchronization supply oscillator was chosen to have a frequency of 4.94197666 MHz, which is one-third of 14.825930 MHz. This oscillator is oven controlled and has a stability of two parts in 10^8 over the required temperature range and an aging rate of less than five parts in 10^{10} per day. In order to get all the harmonics needed for locking the microwave generators and for circuit convenience, the AR6A synchronization oscillator frequency is divided by 16 to obtain 0.30887354 MHz. This latter frequency is distributed to the phase-locked loops associated with microwave generators and the shift oscillators. The phase-locked loops contain circuitry that will hold the frequency at its last value and prevent a frequency jump if the synchronization signal is lost. The AR6A synchronization oscillator supplies the synchronization for all 16 bays of a fully equipped repeater station and is protected by automatic switching to a second unit.

4.3 Signal recovery

The customer-to-customer frequency offset objective of the Bell System is ±2 Hz. This means that carriers used in translating baseband signals up and down in frequency must be very accurate in frequency. This is accomplished by locking these carriers to a primary frequency supply (PFS) which itself is locked to the Bell System master oscillator in Hillsboro, Missouri. Frequency-division multiplex systems such as LMX, MMX, and JMX all operate in this manner. A newly developed OMFS is provided in AR6A terminal stations to meet the frequency offset objective.

In the AR6A System no carrier is transmitted along with the AM sidebands, so one must be accurately generated in the receiving MMGT-R. The IF signal into the receiving MMGT-R can be a few hundred hertz from nominal because of accumulated frequency errors from intermediate repeaters. The receiving carrier generated in the MMGT-R receiver is controlled in frequency to compensate for the accumulated frequency errors by using recovery pilots (one in each half of the IF spectrum). These recovery pilots at the input to the receiving MMGT-R have the same accumulated frequency offset as the rest of the IF signal. The receiving MMGT-R translates each half of the IF spectrum down to the MMG spectrum independently. In each case, the received recovery pilot is selected from the MMG spectrum and is compared in a phase-locked loop with an oscillator locked to the local OMFS. The output error signal from the comparator is used to adjust the frequency of the carrier used in the receiving demodulator. In this way, the receiving carrier "tracks" the frequency offset of the incoming signal and the MMG signal is recovered with no frequency offset.

4.4 Equalization

4.4.1 Static

The objective for amplitude deviation from flatness over the 30-MHz channel for each switch section is ±0.5 dB. To help achieve this objective on a radio hop, each TR bay receiver includes a basic bay equalizer, which compensates for known systematic amplitude misalignments of the microwave and IF components. Mop-up equalizers are available in ±0.5 dB slope and parabolic shapes to enable each switch section to meet its ±0.5 dB objective. The amount of mop-up equalization required is specified automatically from switch section amplitude response measurements made by the TSS-R System. Places are available in the 500A protection switch and in each radio receiver to mount the needed mop-up equalizers.

4.4.2 Dynamic

The objective for time-variant amplitude variations is that they be less than ±2 dB for 99.9 percent of the time. The primary source of transient amplitude variations is selective fading of the microwave signal due to unusual atmospheric conditions. An extensive measurement and analysis program was started in 1970 to determine the effects of selective fading on a 6-GHz microwave radio channel.[5] As a result of computer simulation studies made with the collected fading data, the following three strategies were recommended to minimize selective fading effects:[6]

1. Use space-diversity reception for each hop that has significant selective fading.
2. Use a simple automatic gain control (AGC) amplifier in each receiver.
3. Use a dynamic equalizer that will provide slope and parabolic shape correction at the end of each switch section.

The use of space diversity is recommended to reduce the amount of time the microwave signal is in a deep fade where the largest amplitude variations occur. This is discussed further in Section 4.7. The use of an AGC amplifier in each receiver to compensate for the flat gain portion of microwave signal variations is standard in radio systems. In the AR6A System, the AGC amplifier gain is controlled by the voltage average of the three received radio pilots. To compensate for the shape portion of the amplitude variations, a dynamic equalizer has been designed[7] and is used in each main station receiver. The dynamic equalizer contains one bump network located at each end of the IF band. The gains of these networks are continuously controlled by the voltage difference between the edge radio pilots and the center radio pilot of the received channel. Figure 6 shows a typical fade shape across the IF band before and after equalization.

4.5 Interference

4.5.1 Impairments

An important aspect of any new radio system design is the consideration of the interference environment. The AR6A System has to coexist with the existing FM radio systems, other AR6A routes, and self-interference. One effect of interference is to add to the background noise of the channel; this is compensated for by the inclusion of 32.3 dBrnc0 for interference in the system noise allocation.

Other effects are crosstalk, babble, and tone interference, which can

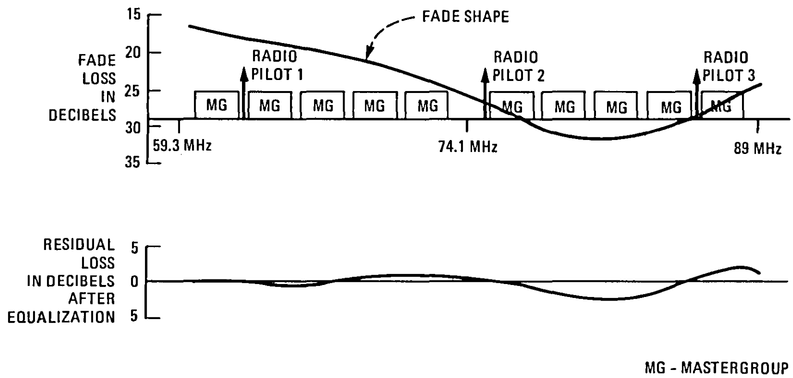


Fig. 6—Dynamic equalization of a fade.

occur when two AR6A channels carrying different traffic interfere with each other.

If both the interfering signal and the desired signal are voice signals and are exactly aligned in frequency, the effect is crosstalk. If the two signals are increasingly offset in frequency, the interference becomes less intelligible until, at ±500 Hz, the interference is judged to be unintelligible crosstalk and referred to as babble. Should the interfering signal be an idle circuit tone or a data signal, the effect will be tone interference regardless of any frequency offset. This type of coupling between two AR6A channels can occur at junction stations or between two routes not sharing a station but in the same geographical area. These effects must be evaluated in terms of the crosstalk, babble, and tone interference objectives. In FM systems, the poorer RF frequency stability and the effects of carrier spreading render these impairments negligible.

Another interference effect is talker echo that results from coupling—primarily antenna coupling—between the east- and westbound channels of the same route. For administrative purposes, the east- and westbound halves of the 4-kHz voice circuits are in corresponding locations in their respective channels. Thus, if these two channels are within 500 Hz of each other, the result is talker echo. Beyond 500 Hz, the effect will be a babbled talker echo, which is judged to be almost as annoying. The magnitude of the talker echo may be described in terms of the effective return loss of the system.

4.5.2 Cochannel interference

To avoid interference from cochannel FM carriers, the center of the AR6A channel is left vacant. The noise effect of the FM sidebands is controlled by properly selecting the channel center gap and frogging

Table I—Impairment calculation summary

Impairment	Objective	Estimated Performance (Percent)
Crosstalk	≤ 1 percent	0.12
Babble/Noise	≥ −10 dB no more than 1 percent of time	0.32
Tone/Noise	≥ −10 dB no more than 1 percent of time	0.55

the mastergroups. The noise effect of another cochannel AR6A System is controlled by requiring at least a 64-dB signal-to-interference ratio.

Studies were conducted to evaluate the ability of the AR6A System to meet the crosstalk, babble, and tone objectives. These impairments are largely controlled by the antenna system discrimination against interfering signals and route layout. Most of AR6A installation would be on existing routes, which means antennas and routes are already fixed. A design choice that was open was the frequency stability of the radio line. As noted in Section 4.2, the microwave carrier supplies required stabilization to keep the radio pilots within the passband of their pick-off filters. Several stabilization plans with different degrees of stabilization were analyzed to determine their effect on crosstalk, babble, and tone interference. These included a plan that introduced a series of fixed frequency offsets on different radio channels to reduce crosstalk. The results showed that a highly stabilized plan was the best for balancing the impairments. A summary of these impairments for the 4000-mile system model is shown in Table I.

One remaining cochannel interference that required analysis was the same route interference that gives rise to talker echo. The requirement for a 4000-mile system is an effective return loss of at least 30 dB. Calculations estimate that the return loss of the radio line is 41 dB.

4.5.3 Adjacent-channel and tertiary inference

The AR6A System has no significant signal-related energy outside of its channel bandwidth. Thus, adjacent-channel noise effects are negligible for adjacent AR6A channels. However, mutual problems occur when an AR6A and an FM channel are adjacent. If the FM channel is an RF squelch-equipped TH-3 System, this feature must be removed because the initiator noise slot is located inside the AR6A channel in a mastergroup position. Without removal the initiator would be continuously operated. Another problem is generated due to the wide bandpass of the IF filters needed in the FM systems to reduce their own channel noise. These filters couple into the FM receiver a portion of the adjacent AR6A signal, which is limited primarily by the cross-polarization discrimination of the antenna system. The noise caused by this interference restricts the number of exposures to 10 or 15. In addition, the tertiary interference effect of a portion of the AR6A signal coupled into the adjacent FM channel is also undesirable. This AR6A signal when coupled back into the AR6A channel will produce modulation on the edge radio pilot and an amplitude shape deviation, depending upon the phase relationship between the signal and the interference. Estimates of these effects indicate they should be avoided if possible, but if they cannot be avoided during route conversion from FM to AR6A, then the AR6A channel adjacent to an FM channel must be either the protection channel or a working channel with the five adjacent mastergroups vacant.

4.6 Linearity

4.6.1 General

Low intermodulation distortion (IM) properly optimized relative to other noise contributors is a basic consideration in the design of a transmission system. It is of particular interest in the AR6A System since the achievement of satisfactory linearity in the transmitting microwave amplifier is a key factor in making the SSBAM Microwave System possible. Transmitting amplifiers with power-handling capacities comparable to those commonly used on current FM systems fall short of meeting SSBAM linearity requirements by some 20 to 30 dB. Achieving stable distortion reductions of this magnitude was thus a crucial design task. Furthermore, although transmitter nonlinearity is dominant, care must be exercised in the design of all other in-line elements to ensure that their cumulative distortion contribution is not excessive.

Since at microwave frequencies and even at intermediate frequencies we are dealing with narrowband channels, the only significant distortion contributors are the odd-order, nonlinear terms. In AR6A, third-order distortion is dominant, fifth order is on the fringes of importance, while seventh and higher odd-order terms are negligible.

A related and crucial design consideration is the question of how the distortions from tandem nonlinear networks combine within and between repeaters. This is referred to as the Law-Of-Addition (LOA) question. Differences between the extreme assumptions of coherent and noncoherent addition strongly influence a variety of design choices. For example, the LOA influences the optimum system drive level, it affects how frequently mastergroups must be relocated or frequency frogged, and it affects equalization strategies. For these reasons, studies and estimates of the LOA received early attention, while subsequent verification of the assumptions was an important element in the field evaluation.

4.6.2 Distortion estimates

To make system distortion estimates, first the distortion contribution must be estimated from individual system elements. This, in turn, must be based on some suitable measured parameter(s) indicative of the unit's nonlinearity. For AR6A computations, a practical approach starts from a power series description of the unit's input/output relationship. Functional ratios of the power series coefficients can then be related to measurable modulation coefficients and vice versa.

To estimate distortion, the broadband message load can be simulated by zero-mean Gaussian noise with appropriate spectral shaping. Multiple convolutions weighted according to power series coefficients may then be used to determine expected IM distortion. In AR6A, since the signal spectrum is flat, the IM noise computation can be greatly simplified by using the work of Y. L. Kuo.[8] Distortion-to-signal ratios are evaluated in terms of power series coefficients, which are then related to measurable modulation coefficients. The final result is expressed in the following relationships, which assume that the odd-order power series has no significant terms beyond the fifth

$$\begin{aligned} N_3 &= 4.23 + 2P_{\text{out}} + M_{\alpha+\beta-\gamma} \\ N_{35} &= 4.77 + 3P_{\text{out}} + \frac{1}{2}\left(M_{\alpha+\beta-\gamma} + M_{2\alpha-2\beta+\gamma}\right) \\ N_5 &= 8.32 + 4P_{\text{out}} + M_{2\alpha-2\beta+\gamma}, \end{aligned} \tag{1}$$

where

N_3 = ratio of third-order distortion power in a narrowband ΔW to signal power in the same bandwidth. The ratio is expressed in decibels and the band ΔW is located at the center of the simulated, spectrally flat, message load,
N_5 = the corresponding fifth-order power ratio,
N_{35} = the corresponding fourth-order power ratio being an interaction term between the third- and fifth-order distortion mechanisms,
$M_{\alpha+\beta-\gamma}$ = a third-order modulation coefficient expressed in decibels,
$M_{2\alpha-2\beta+\gamma}$ = a fifth-order modulation coefficient expressed in decibels, and
P_{out} = total simulated message-load power in dBm at the device output.

The noise-to-signal power ratios can be converted to dBrncOs as follows:

$$\textit{noise}\ (\text{dBrnc0}) = N + P_1 + 86.8, \tag{2}$$

where N is the noise-to-signal power ratio in dB and P_1 is the average signal power per 4-kHz message circuit at 0 transmission level (TL). These relationships involve a number of simplifying assumptions and should be used with caution. For example, the M values are assumed to be constants independent of drive level and frequency; addition of $M_{\alpha+\beta-\gamma}$ and $M_{2\alpha-2\beta+\gamma}$ in the expression for N_{35} is subject to some uncertainty because of the absence of product phase information in the definition of M; and, finally, the expressions apply at the band center of an idealized simulation of the message load. In AR6A applications, cross-checks have been made against pseudorandom noise load measurements[9] with satisfactory agreement if nominal drive levels are not exceeded by more than a few decibels.

With typical values for an output traveling-wave-tube amplifier of $M_{\alpha+\beta-\gamma} = -90$ dB, $P_{out} = +25$ dBm, and $P_1 = -19.6$ dBm0, the estimated IM noise is 23 dBrnc0 per repeater. Taking account of product addition laws, thermal noise contributions, etc., led to the earlier assertion that distortion reductions of at least 20 dB are needed to make the SSBAM System viable.

4.6.3 Law of addition

For a string of n identical nonlinear devices, the law of addition (K) for intermodulation noise (IM) is defined by

$$W_n = K \log n + W \text{ dBrnc0}, \tag{3}$$

where

$$W_n = \text{intermodulation noise of the string,}$$

$$W = \text{intermodulation noise of one device.}$$

Odd-order distortion from some elements of a repeater are going to add systematically corresponding to $K \approx 20$. Other sources of repeater IM are going to accumulate in a random fashion corresponding to $K \approx 10$. The overall law of addition is going to be somewhere between those extremes and will depend on the relative magnitude of the two types of contributors. Clearly, the net value of K is going to be an extremely important factor in estimating total system noise on a long route.

In the case of IM contributed by repeater elements subject to predistortion correction, the residual distortion can be expected to have a random phase orientation. However, some departures from randomness will be experienced if there are systematic degradations in predistorter improvement due to frequency response or aging effects.

Nonlinear elements not subject to predistorter compensation, such as IF amplifiers, will generate IM that adds systematically. Again,

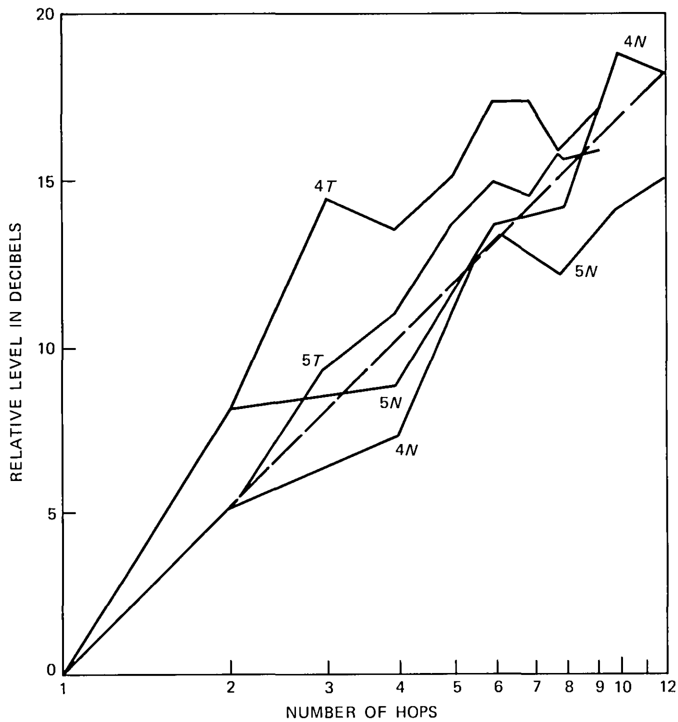


Fig. 7—Intermodulation noise law of addition.

however, there are modifying circumstances such as repeater delay distortion that in this case can reduce the rate of accumulation. Computations involving multiple convolutions of the signal indicate that on AR6A, delay distortion reduces the law of addition from 20 to approximately 19 near band center with larger reductions near the band edges.

For design purposes, law-of-addition estimates were based in broad terms on the power addition of transmitter IM, the voltage addition of receiver IM, and power addition of the combination. Depending on detailed assumptions, this led to estimates for K in the range 16 to 18.

Measurements using pseudorandom noise loading[9] for 12 pairs of repeaters in the initial field installation yielded an average value $K = 16.9$. These same repeaters were then connected in loops of 2 through 12 hops, and the pseudorandom noise load results for radio channels 4 and 5 ($4N$, $5N$) shown in Fig. 7 were obtained. The dotted line in the figure refers to the expected law of addition ($K = 16.9$). Measurements using a three-tone test ($4T$, $5T$) are also shown in the figure. Reasonable agreement with the expected law of addition is observed.

4.7 Application of frequency- and space-diversity protection

4.7.1 Microwave propagation of the AR6A signal

In the early development of AR6A, it was recognized that it would be necessary to characterize the effects of selective fading on the broadband SSBAM signal. Propagation studies were conducted in 1971 on a 26.4-mile path from Palmetto to Atlanta, Georgia. The studies quantified the fading effects on a statistical basis and determined many of the basic system design features, including the choice of the three radio pilots to control system gain, the type and amount of dynamic equalization, and the algorithms for space- and frequency-diversity switching.[10]

In microwave radio system design, multipath (Rayleigh) fading is usually characterized by the equation[11]

$$T(L) = rT_0L^2,\ L < 0.1, \tag{4}$$

where

L = the ratio of received carrier amplitude to nominal or free-space received carrier amplitude,
T_0 = time base or total time of the study,
r = fade occurrence factor representing the fraction of time that selective fading occurs, and
$T(L)$ = time in seconds during the period T_0 that the received carrier amplitude ratio is below the level L.

This relation is descriptive of any single frequency in the AR6A spectrum. The propagation studies expanded this signal characterization to four frequencies. Statistics representative of the fading of any independent frequency in the spectrum were generated with conditioning on the state of the three radio pilots. The equation modeling the fading becomes

$$T(L) = rT_0L^2(U(L) + P(L)), \tag{5}$$

where

$U(L)$ = the conditional probability that the radio pilots are less faded than the ratio L given that the frequency of interest is faded at least to the ratio L, and
$P(L)$ = $1 - U(L)$.

The probability $U(L)$, shown as the curve $L_F = L$ in Fig. 8, is less than 0.1 for fades less than about 36 dB ($L = 0.016$) and increases monotonically with fade depth to about 0.5 for fades greater than 48 dB ($L = 0.004$). Since the radio pilots are used to control the system gain and switching functions, this characterization allows various equalization and switching plans to be evaluated.

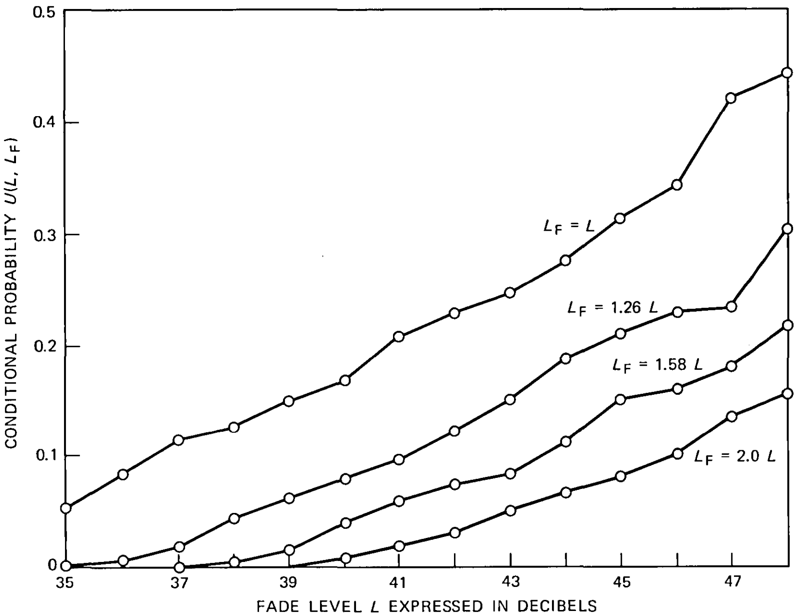


Fig. 8—The conditional probability $U(L, L_F)$.

In addition to the four frequency fading characterization, the 1971 propagation data were also used to characterize the amplitude response of the signal during fading. The data were used as the input to a computer program that simulated the response of the dynamic equalizer. The results of this study are shown in Fig. 9. This plot indicates the expected average severity of gain deviations during fading. The averaging process includes an average over all voice-circuit locations in the signal spectrum. The plot is similar to a cumulative distribution with the maximum fade level of the radio pilots as the independent variable. The 2-dB gain deviation limit corresponds to the dynamic equalization objective. As indicated in the figure, fades with radio pilot attenuation less than 30 dB contribute essentially no gain deviation time. For fades with radio pilot attenuation exceeding 40 dB, comparison of Fig. 9 with the Rayleigh model of eq. (4) shows that almost all of the fade time has gain deviations exceeding ±2 dB. These results led to the recognition that in order to meet the system 2-dB gain deviation objective, space-diversity switching at radio pilot fade levels in the range of 30 to 40 dB would be required.

4.7.2 Frequency- and space-diversity switching during multipath fading

With the recognition that both frequency- and space-diversity switching would be required for AR6A, a fading model including the

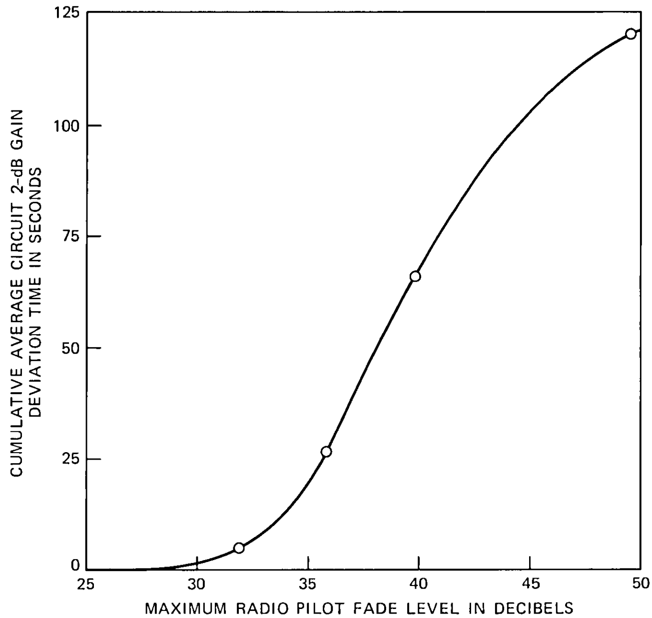


Fig. 9—Average gain deviation time of an AR6A circuit for 600 seconds of multipath fading at the 30-dB level.

effects of both switching systems was developed.[6] The resulting equation representing the fading performance of a hop equipped with both types of switching is

$$T_p(L) = \frac{rT_0L^2}{I_s(L_s)}\left[U(L, L_F) + \frac{P(L, L_F)}{I_F(L_F)}\right], \; L \leqslant L_F < L_S < 0.1, \qquad (6)$$

where

L_F = the pilot fade level corresponding to the frequency-diversity switch threshold,

L_S = the pilot fade level corresponding to the space-diversity switch threshold,

$I_F(L_F)$ = the improvement factor resulting from frequency-diversity switching, typically $I_F > 1$,

$I_S(L_S)$ = the improvement factor resulting from space-diversity switching, typically $I_S \gg 1$,

$U(L, L_F)$ = the conditional probability that the three radio pilots are less faded than L_F given that the frequency of interest is faded to level L,

$P(L, L_F)$ = $1 - U(L, L_F)$, and

$T_p(L)$ = time that the protected signal is faded below level L.

The quantities I_F and I_S vary as L^2 but differ from the conventional,[10] single-frequency improvement factors. Typical values in the current system design are $I_F = 15$ and $I_S = 100$. The conditional probability function $U(L, L_F)$ is shown in Fig. 8 with the relationship between L and L_F as a parameter. The results in the figure are calculated from the 1971 data without smoothing. Since $I_F(L_F) > 1$, eq. (6) shows that higher values of $P(L, L_F)$ are desirable. The increase in $P(L, L_F)$ corresponds to an increase in the fade time that is protected by frequency-diversity switching and a decrease in outage time.

The 2-dB gain deviation time for a path protected by frequency- and space-diversity switching is modeled by the equation[6]

$$T_2(L) = \frac{rT_0}{0.6}\left[D(L_s) + \frac{D(L_F) - D(L_s)}{I_s(L_s)} + \frac{D(L) - D(L_F)}{I_s(L_s)I_F(L_F)}\right], \quad L < L_F < L_s, \qquad (7)$$

where the function D is the 2-dB gain deviation time given in Fig. 9. Equations (6) and (7) formed the basis for evaluating various space- and frequency-diversity switching algorithms.

4.7.3 Reliability and dynamic equalization requirements

During the development process, the performance of the radio system with respect to the reliability and dynamic equalization objectives was evaluated for several space- and frequency-diversity switching algorithms. In particular, various combinations of the switching levels L_S and L_F were studied. System performance for both voice and voiceband data was included. It was found that lower values of L_S and L_F improved the reliability performance but degraded the dynamic equalization. The current system design represents an approximate optimum in the trade-off between these requirements.

The space-diversity switching level was chosen to correspond to a 36-dB fade on a 27.1-mile path. For longer or shorter paths, the switching level varies inversely with the received signal level. The frequency-diversity switching level was chosen to be 53 dBrnc0 corresponding to a 44-dB fade on a 27.1-mile path. With this switching arrangement, approximately one-half of the hops in an average fading area are expected to require space diversity.

4.7.4 Computerized implementation

An interactive computer program was developed to facilitate the application of frequency- and space-diversity protection to new AR6A switch sections.[6] The evaluation of a proposed section's performance with respect to the outage and dynamic equalization objectives is accomplished by the program after input data are supplied by the program user. The data are used to characterize the expected fading on the various hops in the section. For each hop, the input data include a climatic factor; the mean annual temperature; the hop length, roughness, and fade margin; and a description of any proposed space-diversity antenna.[10] The program first evaluates the performance of the section without space diversity. The estimated performance is compared to the objectives after prorating them to the length in miles of the section. If the objectives are met, the program stops and the route can be installed without space diversity. If they are not met, the program reevaluates the performance with space diversity applied initially on the worst fading hop. The program continues in this manner until the objectives are met. The route would then be installed with space diversity on the specified hops.

4.8 Mastergroup frogging

Noise, reliability, and dynamic equalization in AR6A are functions of voice-circuit location in the IF passband. Circuits located near the IF band center receive relatively high levels of interference noise from FM-FDM and FM-TV systems. Reliability and dynamic equalization are best for circuits located near one of the radio pilots. Because of these variations, more uniform performance is attained on long trunks by varying their location within the IF spectrum. The variation is accomplished by interchanging or frogging mastergroups at specified distances. The maximum allowable frogging distance chosen for field applications is 800 miles. This value provides acceptable system performance and reduces the amount of multiplex equipment required solely for frogging. The effects of frogging distance on system performance are described in the following sections.

4.8.1 Noise performance

The system noise components that are influenced by frogging include: intermodulation noise, radio pilot modulation noise, and worst-circuit FM RF interference noise. The levels of these components all increase with frogging distance and are listed in Table II for distances of 400, 800, and 1600 miles. The intermodulation noise increase with

Table II—Noise performance with various frogging rules

Noise Type	Noise in dBrnc0 400-Mile Rule	800-Mile Rule	1600-Mile Rule
Intermodulation	31.7	33.3	34.0
Radio pilot modulation	16.1	21.1	26.1
Worst-circuit FM RF interference	27.4	30.4	31.9

frogging distance is due to the radio receiver. The receiver intermodulation products produced on hops within a frogging section are nearly coherent. A small reduction, estimated to be 2.6 dB for an 800-mile frogging section, results from the phase characteristic of the IF filter. The receivers' products in different frogging sections are incoherent because of the changed signal. The radio pilot modulation noise also increases with frogging distance as a result of hop-to-hop coherence.

Although the average noise resulting from the FM RF intereference does not change with frogging distance, the distribution of the noise does. Circuits located near the center radio pilot experience maximum interference. When the frogging distance is increased beyond 400 miles, some circuits receive significantly more noise than the average. The resulting worst-circuit noise is shown in Table II.

4.8.2 Reliability and dynamic equalization performance

Since the noise initiator slots for the 500A Protection Switching System are located near the radio pilots, voice circuits located near the pilots receive relatively good selective fading protection. For these circuits unprotected fade time is small. For circuits located between the radio pilots, some selective fades can cause outage without producing enough noise in the slots to initiate a switch. As a result, the outage performance of the system varies with IF frequency. With a 400-mile frogging rule, all mastergroups transmitted 4000 miles occupy each mastergroup location in the signal spectrum for approximately 400 miles. When the frogging distance is extended beyond this interval, the performance is no longer uniform. With the 800-mile rule, calculations indicate that the outage of the worst circuits will exceed the average circuit by 44 percent. To ensure that these circuits meet the overall multipath fading objective, an algorithm was added to the computer program for the application of space diversity. For switch sections located in frogging sections between 400 and 800 miles in length, worst-circuit performance is calculated and compared to the objective. Additional space-diversity protection is then required for the switch section to meet the system requirement. This procedure allows the engineers responsible for a route to trade off the cost of additional multiplex required for a shorter frogging section with the cost of additional space diversity required for a longer one.

For frogging distances greater than 400 miles, the dynamic equalization performance of the system parallels the reliability performance. Worst-circuit gain deviation time with the 800-mile frogging rule is approximately 48 percent greater than the average circuit time. This factor is also included in the computer program for the application of space diversity, and the worst circuits are required to meet the system objective.

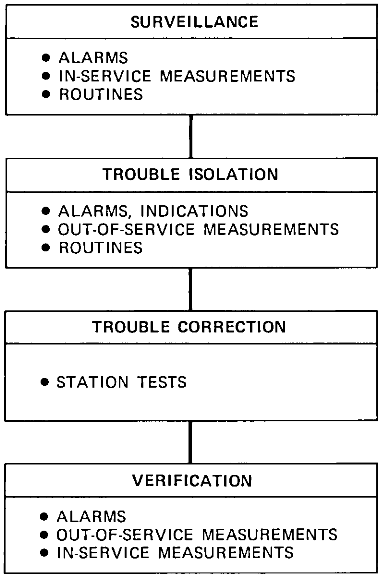


Fig. 10—AR6A generalized maintenance plan.

4.9 Maintenance

4.9.1 General

Maintenance was discussed in general terms in Section 1.6 with particular mention of the importance of adequate transmission surveillance capability. In AR6A, the new surveillance functions are combined with traditional remote alarm and reporting features to form an overall maintenance plan that is summarized in very general terms in Fig. 10.

The alarm reporting system is a full-time monitoring facility using either C1 or E-type alarm remotes under the centralized jurisdiction of a SCOTS or TASC central minicomputer. In general, remote alarm reporting is limited to isolating trouble to a particular station and direction of transmission, not the particular unit in trouble. Locally, however, more detailed visual indications and test points simplify the trouble isolation process. This approach is based on the premise that prior knowledge of the detailed trouble condition is unnecessary if it is assumed that, on a station visit, the technician is equipped to deal with any on-site problem or any additional problems that become known enroute.

Except for a few legally required routine tests, maintenance is performed on an as-needed basis. When a trouble has been identified and isolated, a technician is dispatched to the trouble site with a defined set of spare parts and station test equipment. The principal station test sets are:

1. Transmitter-receiver transmission test set
2. Scanning intermodulation test set
3. Portable rubidium frequency standard
4. Pilot selection test set
5. TWT power supply test load.

4.9.2 Transmission Surveillance System

The AR6A Transmission Surveillance System—Radio (TSS-R) consists of programmable equipment for access and testing located at radio switching stations. It can perform a variety of transmission measurements under the direction of a central minicomputer. Communication between the central computer and remote measuring sites is via *DATAPHONE** data communications service with its attendant flexibility and expandability. Unlike the alarm reporting system, TSS-R measures system performance on a scheduled or as-needed basis, providing snapshot rather than continuous information.

Figure 11 is a block diagram of TSS-R. The central minicomputer has a peripheral disk storage capacity of approximately 20 Mb allocated to program, database, and data storage functions. Automatic call-out units provide DDD access to the measuring remotes as well as to maintenance and other centers programmed to receive reports. It is expected that five central computers will be capable of covering all AR6A deployment within the continental United States.

At switching stations, test access, measuring equipment, and data interchange with the central computer are under the control of a microprocessor with its associated firmware. Physically, these functions are grouped in the TSS-R distribution bay as indicated in Fig. 11. All measurements are performed by a Selective Transmission Measuring Set (STMS) with a frequency range of 10 kHz to 160 MHz. The set is controlled by the distribution bay microprocessor via an IEEE control bus. Programmable functions include level and frequency measurements, bandwidth selection, and the selection of a phase-jitter measuring mode.

Switched access for measurement purposes is provided at transmitting and receiving mastergroup combiners and at transmitting and receiving 500A switches. Measurements at mastergroup combiners can be made on a bridged, in-service basis only; measurements at the 500A switches can be made either on an in-service or out-of-service basis.

In-service measurements can be made conventionally using bridging hybrid transformers in conjunction with director switches to route the selected test point to the STMS. Signals available for measurements

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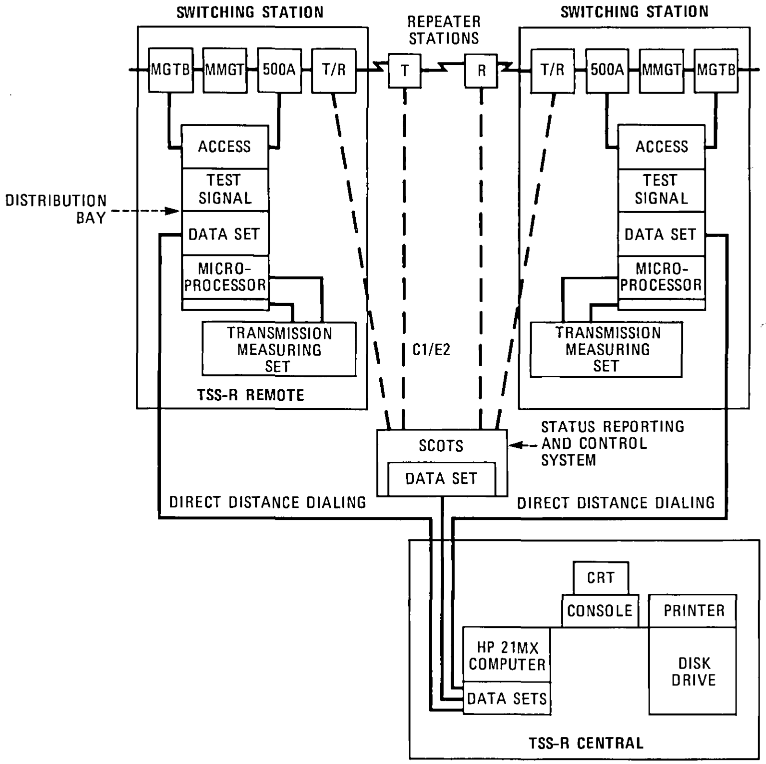

Fig. 11—Transmission Surveillance System.

are elements of the normal system load such as mastergroup pilots, system pilots, and intermastergroup noise. Since in-service measurements are made without affecting the main transmission path, they are made frequently and routinely at intervals of approximately one week.

On the other hand, out-of-service measurements require that service be temporarily transferred to the protection channel. This frees the selected working channel for end-to-end switch section measurements. To measure the channel, access at the head end of the switch section is provided by the 500A maintenance switch as indicated in Fig. 12. Since operation of a maintenance switch opens the channel and removes normal pilots, a substitute set of pilots must be supplied via the maintenance measuring path (see Fig. 12). At the receiving end of the switch section, the channel under test is routed to the STMS for making the selected out-of-service measurements.

Since the operation of a maintenance switch can be done under program control, safeguards must be provided against untimely oper-

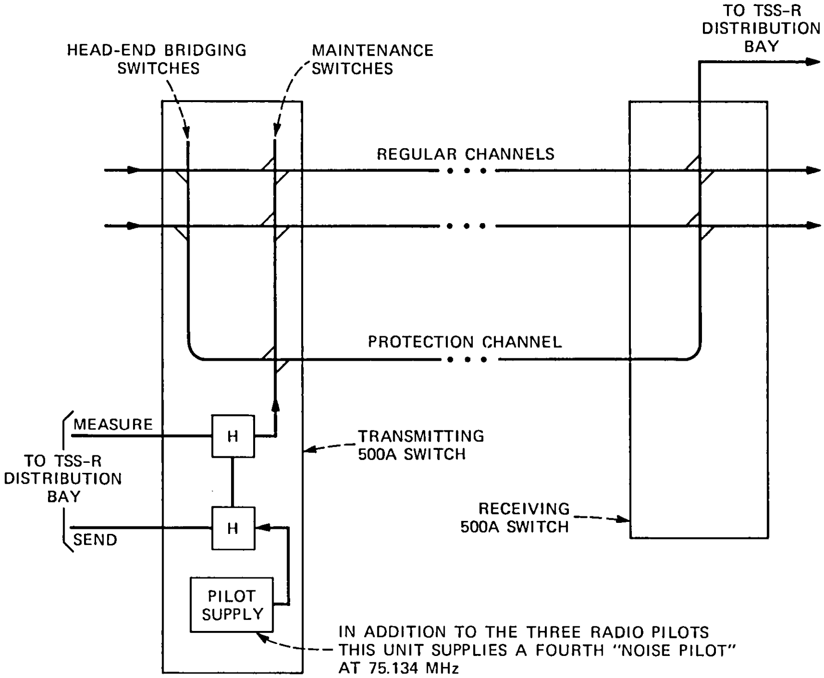

Fig. 12—Simplified diagram of switch section out-of-service test access arrangements.

ation of the switch by TSS-R. This is achieved by making the 500A control circuits the ultimate arbiter of all switching decisions.

Some out-of-service measurements require only that the head-end of the channel under test be terminated. Other tests require the insertion and measurement of a test signal at the head end of the channel as well as its measurement at the receive end. Provisions for inserting test signals, combining them with pilots, and providing head-end measuring access are shown in Fig. 12. For example, to measure out-of-service amplitude response, a comb signal with a tone spacing of 130 kHz that encompasses the entire IF band is available from the distribution bay. This signal, when needed, is sent to the test signal input port (Fig. 12) and measured at each end of the channel under test. The measured data are returned to the TSS-R central minicomputer where a comparison of transmitting and receiving levels provides the desired amplitude response characteristic for the channel.

Special mention is made of the following out-of-service switch-section measuring capabilities:

1. Spurious modulation—This involves the insertion of a +5 dBm0 test tone (83.168 MHz) at the head end of the switch section. At the receiving end, TSS-R looks for spurious modulation sidebands on the

test tone. These might be due to modulation of the test tone by spurious switching tones from a bad power supply, spurious modulation by a faulty microwave local oscillator, or spurious modulation by some other unwanted mechanism.

2. Linearity—This measurement is of special interest for verifying continued satisfactory adjustment of transmitter predistorters. Two high-level (+17 dBm0) test tones located ±44 kHz on either side of the 74.13-MHz channel center frequency are inserted at the head end of the switch section. Third-order nonlinearities produce IM products located 88 kHz on either side of the radio-line pilots. These products are measured at the receiving end of the switch section as an indication of switch section linearity. Positioning the +17 dBm0 full-load test tones close to band center minimizes the impact of cochannel interference on other systems. In other words, the tones are no worse from an interference standpoint than cochannel FM carriers.

3. Transmitter gain check—The +17 dBm0 test tones mentioned in (2) can also be used as a coarse check on transmitter gain. When turned on at the head end of a switch section, power monitors on the output of each succeeding transmitter provide a remote indication if the power is within a ±3 dB window of nominal.

A complete listing of all in-service and out-of-service tests is given below.

1. In-service tests—switch section
   (a) Pilots—level and frequency
   (b) Intermastergroup noise
   (c) Pilot slot noise—before pilot insertion
2. Out-of-service tests—switch section and hop by hop
   (a) Idle noise
   (b) Noise and tone scan
   (c) Linearity
   (d) Amplitude response
   (e) Interference
   (f) Cross-polarization discrimination
   (g) Transmitter gain
   (h) Space-diversity switch point
3. Other tests
   (a) Head-end pilot supply on 500A
   (b) Phase jitter—in-service and out-of-service
   (c) Diagnostic tests

4.9.3 Stress-test and trouble-isolation features

Distinct from, but enhanced by, TSS-R are a number of supplementary features for stress testing and trouble isolation. Central to implementing these features is the capability to remotely operate any selected resupply switch at intermediate repeaters via the alarm reporting and command system. When a resupply switch is operated, a locally generated set of pilots is inserted into the channel under test. By measuring the channel with TSS-R at the receiving end of the switch section, it is possible to characterize systematically shortened portions of the switch section to isolate trouble.

To ensure that a resupply switch can only be operated remotely on an out-of-service channel, the switch command is enabled by the presence of the noise pilot on the out-of-service channel (see Fig. 12). Since the noise pilot can only be present subsequent to the operation of a head-end maintenance switch and the prior transfer of service to a protection channel, the requisite protection is provided.

An important additional function of the noise pilot is associated with the switching system preemption feature. If TSS-R is measuring a foreshortened portion of a channel in a switch section, rapid restoral of the intermediate resupply switch is essential if the channel under test has to be preempted for service. Since a first step in the preemption process is to pull down the head-end maintenance switch and thereby remove the noise pilot provided by the head-end resupply, all downstream resupply switches are automatically disabled and returned to normal. This avoids the slow restoral procedure that would result if resupply switches had to be reset via the alarm reporting and command system.

Trouble isolation capabilities associated with the remote control of resupply switches are further enhanced by providing means to remotely modify the resupply signal. For example, the two high-level test tones used for switch section linearity measurements can also be turned on at intermediate resupply points. This provides the capability to isolate hops with high nonlinearity by looking at systematically foreshortened portions of the switch section. The isolation of linearity problems to a particular hop is further enhanced by turning on the aforementioned linearity test signal at a selected intermediate station; but instead of applying the radio pilots at their normal −10 dBm0 level, they are applied at 0 dBm0. This enhances the third-order nonlinearity product contributed by the first hop of the foreshortened switch section by 10 dB. At succeeding receivers on the switch section, AGC restores the radio pilots to −10 dBm0 and correspondingly reduces the high-level test tones by 10 dB to +7 dBm0. Thus, the third-order products contributed by succeeding hops are reduced by 20 dB, thereby making the first hop contribution dominant.

Another stress test feature is provided by remotely commanding the resupply pilots to be inserted 32 dB below their normal level of −10 dBm0. When this is done, AGC on the first receiver beyond the resupply point increases its gain by 32 dB to restore the pilots to their normal −10 dBm0 level at the receiver output. This enhances the front-end idle noise contribution of that particular receiver by a corresponding amount and will, therefore, be dominant when TSS-R measures noise at the end of the switch section. In effect, therefore, this measurement gives an indication of the receiver noise figure on the artificially faded hop. There is a corresponding 32-dB enhancement of incident interference on the faded hop providing a means for identifying the interference level and point of entry. This same technique may also be used to measure the level of the nearest radio pilot in the adjacent AR6A channel to obtain a measure of cross-polarization discrimination on a hop-by-hop basis.

V. TESTING

5.1 Laboratory testing

A test facility was established at Merrimack Valley to perform initial testing of AR6A bays. This test facility served the needs of developing shop and field testing information, initial testing for radio bay components, and pursuing problems observed in the later planned field trial program. Two sets of frequency-compatible bays were installed along with a support bay. The bays were configured so that each set had a main-station-configured bay and a repeater-station-configured bay. This arrangement allowed for the greatest flexibility for an ongoing testing program.

Figure 13 depicts the laboratory bay arrangement. Bays 1 and 2 were configured as main station bays. The waveguide runs connecting repeater bays 3 and 4 contain RF preamplifiers and the loss of each run was built out to simulate a nominal hop length of 27.1 miles (61.7 dB of section loss). Two MCSS supplies were available in the support

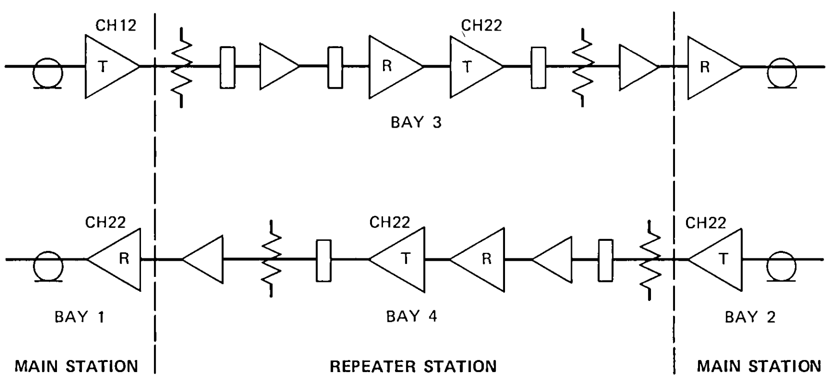

Fig. 13—Laboratory bay arrangement.

bay to allow for all bays to be locked to the same unit or for separate MCSS signals for bays 1 and 2 as testing required.

Initial testing concentrated on tests and measurements to ensure that an assembled bay met system requirements. Extensive tests were performed to isolate and reduce noise pickup and cross-coupling of signals in the bay wiring. These and results from the later field trial showed that a frame filter was necessary to filter the power from the −24 volt plant if noise objectives were to be met.

Modular design of the radio bay implied that a set of pretested modules could be inserted into a basic framework and a working bay obtained with a minimum of alignment. Initial testing was directed toward this goal. It was necessary to specify a series of tests for each module to ensure that when a set of modules was assembled into a bay, the bay performed satisfactorily. Test procedures for the overall bay were then developed for bay alignment. Once these procedures were deemed sufficient, they were incorporated into Bell System Practices (BSPs), which were then debugged using the laboratory bay facility.

5.2 String tests

Since all components of the AR6A System were designed and available at Merrimack Valley, a tandem test arrangement was achieved by utilizing trunking between laboratories to construct a complete system comprised of MGTBs, MMGT-R multiplexing equipment, 500A Switching System, and the radio bays. This permitted injecting a signal into the system at basic mastergroup level and testing for switching transients with tandem units. A fade could be introduced producing a space-diversity switch, and as the fade was increased, correct switching activity for the 500A switch and the pilot resupply function in a radio bay could be observed and tested.

Noise load tests were performed on the tandem system, first on the MMGT-R connected back to back, then with the loopback at the 500A switch, and finally with a radio bay. This allowed testing of up to a four-hop switch section.

Phase jitter was also investigated on the system. Figure 14 depicts the test setup for a typical measurement. Pictures A and B in the figure show examples of the demodulated phase-jitter voltage display on an oscilloscope and a spectrum analyzer, respectively. Peak-to-peak switch-section jitter, measured in this fashion, is typically 1.2 degrees.

This laboratory facility was used constantly over the next two years as the minitrial was conducted in western Massachusetts and the field trial was conducted in Missouri. Problems seen in the field were

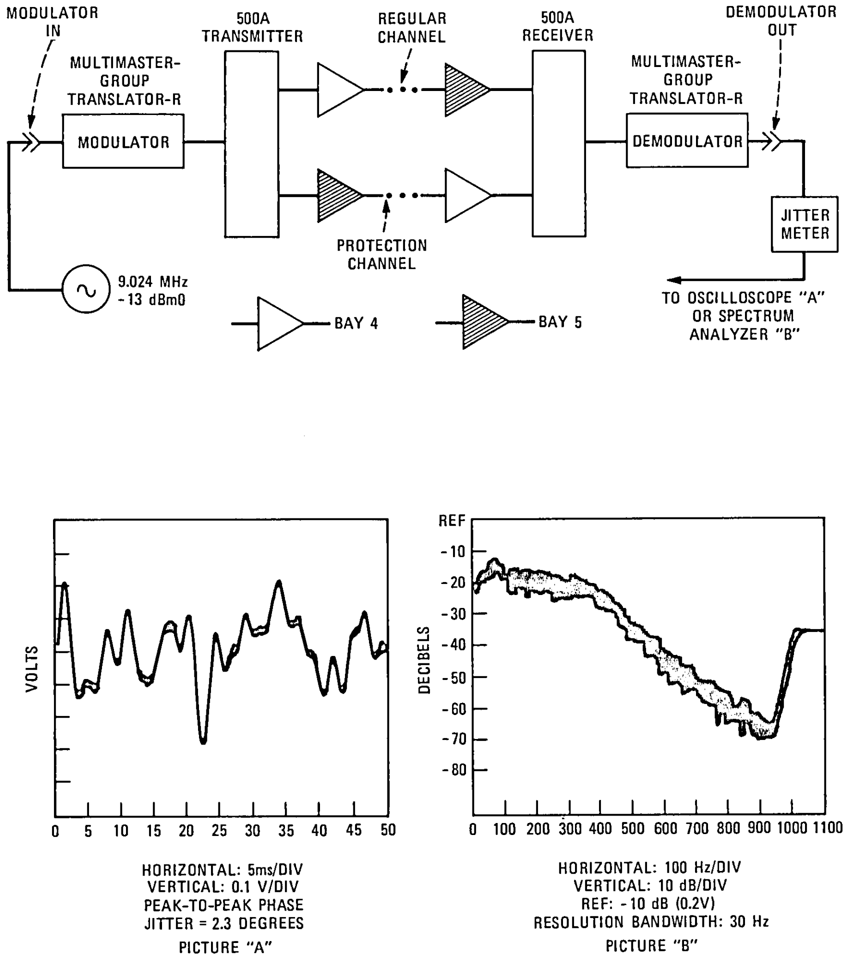

Fig. 14—Phase-jitter measurements.

pursued on the laboratory setup, thus eliminating a great deal of travel time and allowing testing, which led to a rapid resolution of problems.

For example, during the minitrial in western Massachusetts, a fade deep enough to produce either a space-diversity switch or pilot resupply switch caused multiple switching activity. The laboratory setup showed that although the average signal level was insufficient to call for the switch, switching was being initiated by noise spikes due to a peak detector with minimum filtering. Modifications were made to the detector circuitry such that switching was done on an average value over a few milliseconds rather than on an instantaneous peak value, thus eliminating the great number of switches that were occurring.

5.3 Ashburnham-Wendell field trial

5.3.1 Configuration

A field trial consisting of a one-hop loopback of AR6A prototype radio bays was conducted during the period of October 1977 to July 1978 between Northeastern Area Long Lines radio stations in Ashburnham and Wendell, Massachusetts. The Ashburnham station was equipped as a main station and the Wendell station as a repeater site. The hop length was 29.3 miles, and BTL model 656A microwave preamplifiers were used at both receivers. The Ashburnham radio bay transmitted on radio channel 22 (6226.89 MHz) with vertical polarization and the Wendell bay transmitted on channel 12 (5974.85 MHz) with horizontal polarization, which made both bays standard in their channel arrangements. Both stations had prototype support bays that contained the pilot resupply and microwave carrier synchronization supply (MCSS) equipment. In addition to the radio equipment, each station was equipped with a PDP-8 computer which served as a Data Acquisition System (DAS) for complete monitoring and recording of all bay alarm and indication activity. A simplified diagram of the trial equipment is shown in Fig. 15.

5.3.2 Installation and alignment

As a test of the modular design concept Western Electric installers installed the basic bay framework and wiring. The actual assembly of the TR modules in the bay framework was performed by AT&T Long Lines personnel. The installation process was photographed and these photographs along with the documentation from Bell Laboratories were used to generate the installation Bell System Practices.

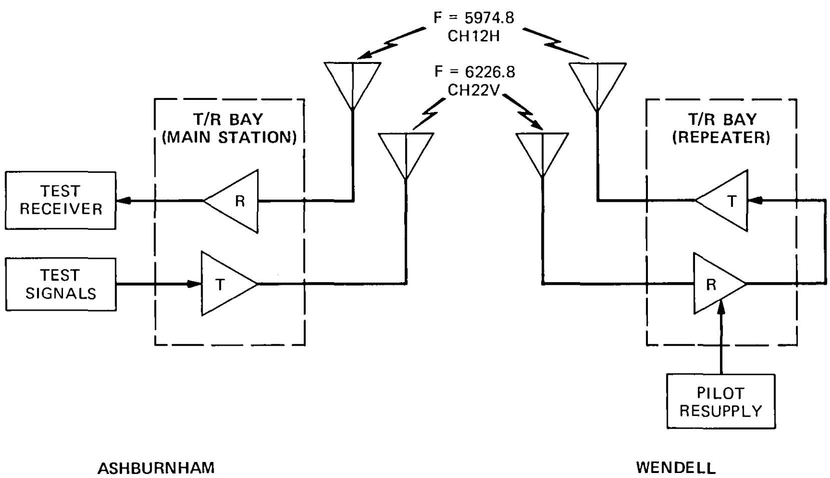

Fig. 15—Ashburnham-Wendell field trial equipment configuration.

5.3.3 System performance

Prior to shipping, the AR6A radio equipment underwent extensive laboratory measurements to fully characterize it. The measurements included thermal and intermodulation (IM) noise, and amplitude response. Initial system measurements made on the Ashburnham-Wendell installation showed good agreement with the laboratory results. Received signal levels agreed with calculated values indicating nominal section loss in the two directions of transmission. These results were further confirmed by thermal noise measurements. Long-term measurements were conducted on received signal strength, IM performance, and thermal noise. The results of these measurements indicated performance adequate or better than that required to meet system requirements. In particular, the long-term behavior of the IM noise was encouraging, indicating that the predistorter realignment intervals would be acceptably long. Examination of the received signal strength and noise slot data during fading activity indicated that the AGC amplifier and dynamic equalizer responded adequately and that space-diversity switch requests were generated at the planned level of 36 dB of fading.

5.4 Factory testing

Historically, radio bays have been installed and tested on site by Western Electric installers. Fully tested and operating bays are then turned over to the operating company. The AR6A modularity concept changed this procedure to one where the customer receives factory tested modules and assembles them to produce a working bay. To ensure that the customer could bring his equipment on line in a timely manner, two procedures were initiated. First, a level of spares was specified that should ensure inhand replacement units to offset initial failures and units damaged in shipment. Secondly, a factory testing program was initiated requiring that all modules be assembled in a test bay and functionally tested before shipment. Both of these procedures are now standard practice.

5.4.1 Bay alignment

The set of modules comprising a radio bay are installed into test bays and then the bay is aligned. Any units not functioning correctly are replaced and trouble sheets on the failed units are filed. The data collected in this procedure are then used to identify problem units and the type of problem often helped identify the solution.

5.4.2 Back-to-back testing

To ensure that a quality product was shipped from the factory, the eight performance tests listed below were performed on each aligned bay:

1. Amplitude response
2. Phase jitter—TR pair
3. Phase jitter—microwave generator
4. Spurious tones
5. Spurious modulation
6. Thermal noise
7. Intermodulation noise
8. Pilot resupply operation.

When factory production was first begun, the bays were produced in frequency-compatible pairs. After bay alignment was completed, the pair was connected back to back (at RF) and the performance tests conducted. As production increased, test bays containing a frequency shifter were constructed so that the performance test could be performed on individual bays. After 126 bays had been tested it was determined that other bay tests were sufficient, so that back-to-back tests were discontinued.

5.5 Missouri trial

During the period from July 1979 through June 1980, an initial evaluation of AR6A was conducted on a six-hop trial switch section between Hillsboro and Windsor, Missouri. This was part of a new 6-GHz route being established between Hillsboro, Missouri, and La Cygne, Kansas, for the first service route of AR6A (see Fig. 16). To allow field evaluation of upper and lower sideband transmitter modulators, radio channel 4 employing the lower sideband and channel 5 employing the upper sideband were installed. This arrangement also allowed for the measurement of adjacent-channel interference, an important factor in overall noise performance.

5.5.1 Installation

The 6-hop trial route was equipped as a two-way, 1×1 frequency-diversity protection system. Hillsboro and Windsor were equipped as terminal main stations and the five stations, Richwood, Rosati, Brinktown, Barnett, and Cole Camp were the connecting repeater stations. Installed at each of the repeater stations were four repeater bays and a support bay. Hillsboro, which is a terminal main station, was equipped with standard radio bays configured for main station application. In addition to the radio bays, a 500A protection switch, MMGT-R, 500B protection switch, and MGTB multiplex were also installed.

Windsor is a repeater station on the final expansion of the route to La Cygne, but for this trial was to serve as a terminal main station for AR6A. This required that two transmitter-only bays and two receiver-only bays (standard options) be installed at Windsor for normal

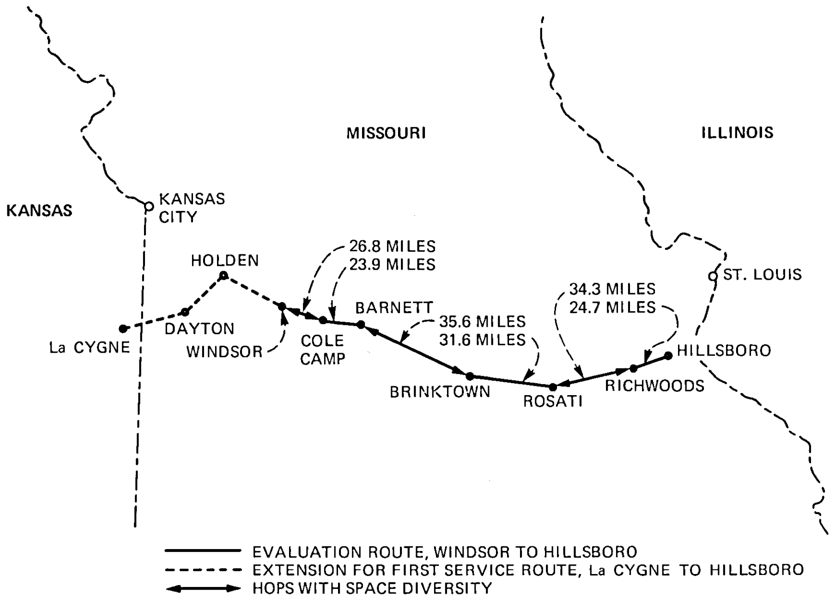

Fig. 16—First service route of the AR6A System.

connection to the indoor waveguide. And, of course, a 500A protection switch, MMGT-R, 500B protection switch, and MGTB multiplex were also installed. To allow signals transmitted from Hillsboro to be looped back to Hillsboro, a set of mastergroup connectors also were installed at Windsor.

Alternate hops on this route were equipped with the space-diversity option. This required that a space-diversity antenna be installed at all stations but Hillsboro.

A Bell Laboratories prototype TSS-R remote distribution bay was installed at both Hillsboro and Windsor. This allowed field evaluation of the distribution bays and their interaction with the TSS-R central minicomputers located at Bell Laboratories.

In addition to the radio equipment, each station was supplied with a PDP-8 computer that served as a Data Acquisition System for complete monitoring and recording of all bay alarm and indication activity. This system was connected to a PDP-8 computer at Bell Laboratories, Merrimack Valley, to allow for continual monitoring of the trial.

5.5.2 *Antenna tests*

The trial route, being part of a new 6-GHz route, had newly installed antennas between Hillsboro and Richwood and newly installed space-diversity antennas. A series of antenna measurements were made to characterize the transmission environment for AR6A radio. The measured parameters included section loss (SL), amplitude response, cross-polarization discrimination (XPD), and front-to-back (FB) ratio. In addition, checks for RF interference were made at each receiving antenna. The test results indicated generally good antenna performance; however, one antenna at Cole Camp did have to be reoriented slightly to meet AR6A System requirements. Junction interference from a TH-3 crossing route at Rosati was severe and had the potential to cause an improper sequence of operation of the space-diversity and 500A switching systems. This situation was corrected before the AR6A hop and switch section tests began.

5.5.3 Hop tests

The installation and alignment of the TR bays were performed by AT&T Long Lines technicians and served as a field trial for the initial issue of the BSPs. Radio bay hop tests were then performed jointly by AT&T Long Lines communication technicians and Bell Laboratories personnel. These tests afforded an opportunity to update and refine the hop turn-up BSP. These tests check such parameters as path loss space-diversity switch point, and pilot resupply initiate switch point.

5.5.4 Switch section tests

Switch section tests were performed in three parts. First, those tests necessary to condition a switch section for service were performed. Second, a series of tests were performed to measure the channel performance, and, third, testing was done with TSS-R to verify its performance. Initial switch section testing normally is performed by TSS-R, as discussed in Section 4.9.2. However, at the start of this field trial, TSS-R was not in operation. The same measurements were performed manually as would have been performed by an operational TSS-R System to condition the switch section for service. Performance of the test in this manner served two purposes. First, a database was generated for comparison with initial TSS-R measurements and, second, the latter measured performance of the switch section would determine if the set of TSS-R tests was adequate to condition a channel for service.

The amplitude response of each of the four channels of the switch section from 500A transmitting switch to 500A receiving switch were measured. The results of these tests were then used to compute the number and type of 989-type mop-up equalizers required to equalize the channel to a 0.5 dB flatness. Figure 17 shows a typical radio channel response after amplitude equalization. These tests showed that ordering information for mop-up equalizers did not supply a proper selection of equalizers. New ordering information for the mop-

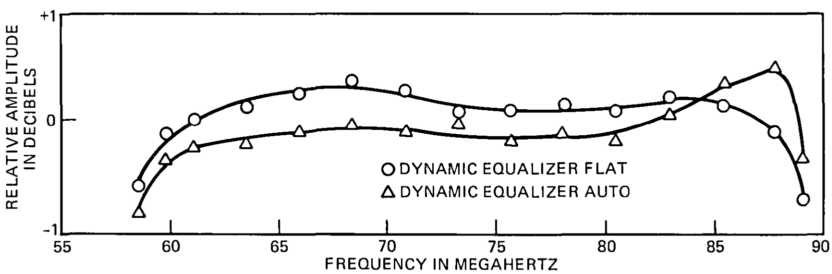

Fig. 17—Typical radio channel response after amplitude equalization.

up equalizers was generated based upon the results of these tests and analysis of factory test data from the initial bays.

The rest of the out-of-service switch section tests listed in Section 4.9.2 were performed on the four channels. The measured thermal noise was slightly higher than desired. This result was not totally unexpected since some of the initial TWTs had a higher noise figure plus gain sum than specified. This sum (noise figure plus gain) is now used as one of the screening tests for tubes used in systems now operating.

A test of system linearity was of significant importance. Hop and switch section linearity were specified based upon the system model. During the trial, it was found that even though every transmitter met its linearity requirement, some hop and switch section limits were difficult to meet. It was determined that these limits were too stringent because of such variables as hop length, systematic level error, and intermodulation noise contribution from the end hop containing a 500A switch. As a result, new limits for hop and switch section linearity were developed that were more realistic, yet ensured system integrity.

The noise and tone scan performed on each channel showed a few undesired tones in the band, which were identified as originating in the multiplex. These tones are harmonics of carriers and, with production-type filters, are eliminated. With the out-of-service tests completed, noise loading tests began. During this second phase of testing, three types of noise load tests were made. The first type was performed using a test signal generated at basic mastergroup level. Ten basic mastergroups of noise were generated and loaded onto the system at the MGTB. The second type of noise loading was performed by bridging on a live signal obtained at the mastergroup distribution frame. This allowed a comparison of random noise load results with a live signal result. The third type of noise loading used a pseudorandom noise source.[9] These tests were performed to determine (1) the level and type of nonlinearities in the system, and (2) the law of addition, i.e., how these nonlinearities add on a hop-to-hop basis.

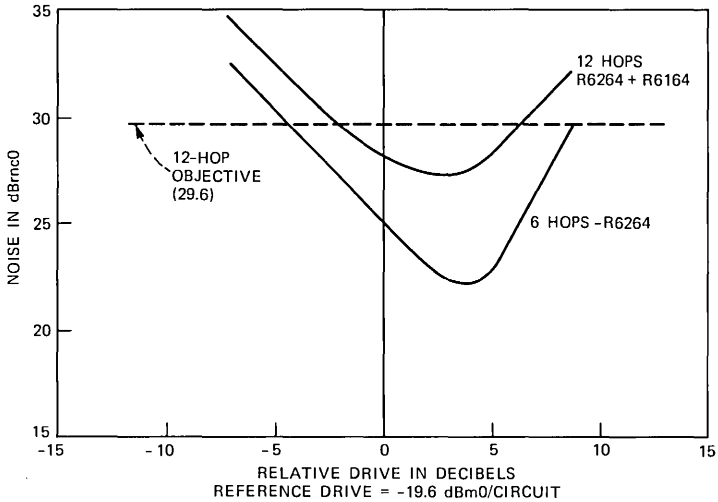

Fig. 18—Noise load tests made during the AR6A field evaluation.

Random noise load measurements were made first on each of the six-hop channels using a random noise load. This measurement was made by looking in a noise slot in each of the ten mastergroups. Figure 18 depicts the results of one such measurement. Measurements were then repeated for a 12-hop switch section by looping back at Windsor first at the 500A, and then at the MGTB employing mastergroup connectors between the transmit and receive MGTB. The results of 6- and 12-hop switch section measurements are shown in Fig. 18. These types of measurements were made early during the field trial, again after several months of operation in which no particular effort was made to keep the channel in peak operating condition. They were then repeated near the end of the trial after all transmitters were realigned to ensure their best linearity performance. A hands-off period of two months was then allowed, and the noise loading tests were repeated to obtain aging information.

Noise loading was also performed using live signal load before and after the hands-off period. The tests employing a live load used nine mastergroups bridged from the mastergroup distributing frame and one mastergroup of random noise containing the measurement slot. This mastergroup could be placed on any of the ten mastergroup positions in the channel. The results showed slightly better performance than the random signal, since the average load with a live signal is less than a noise load due to varying talker activity and volume.

Pseudorandom noise loading allowed measurement of IM distortion produced in the channel. These measurements were performed on the radio portion of the switch section only, that is, from transmitter "in" to receiver "out" of the switch section. The test signal was applied at the transmitter "in" on the radio bay at Hillsboro and recovered from the receiving "out" port of the same bay. To complete the circuit, a patch was made at Windsor to loop back the signals between receiver and transmitter. The receiver of the test set measures the total IM signal (generated by the 12 radio bays) that falls into a narrow (3-kHz) band. These measurements were made for the expected range of system signal powers to ensure that the IM products were within limits for all operating conditions.

As we discussed in Section 4.6.2, the IM products and their law of addition when a large number of repeaters are connected in tandem is of critical importance. A scheme was derived whereby the law of addition could be determined from a series of pseudorandom noise measurements. To perform these measurements, it was necessary to have personnel present at three stations to perform the patching so that each repeater could be measured individually and then in combination. The results of the measurements were discussed in Section 4.6.2 with the results shown in Fig. 7.

A period of the trial was devoted to TSS-R testing. During this phase of testing, each of the TSS-R measurements was performed remotely from the Merrimack Valley central computer and then repeated manually on site. This was to ensure that both the proper measurement was being made, that it was being made on the intended unit and, finally, that the results were correct. Maintenance of the AR6A System relies upon a combination of SCOTS alarm and radio indications and TSS-R in-service and out-of-service measurements. When a problem occurred on a channel, the indications from SCOTS in conjunction with TSS-R measurements were used to identify the problem and its location, so that maintenance personnel could be dispatched to that location with the necessary equipment and spare parts to effect a repair. As a result of this brief evaluation period, the maintenance scheme was judged viable, but further documentation was necessary for the users to take full advantage of the powerful tools put at their disposal.

During the course of the field trial, several minor problems were uncovered in the functional operation of both radio bays and associated 500A Switching System and the multiplex. Identification of these problems allowed for the necessary circuit changes to be implemented into the product prior to full production.

VI. HOT STANDBY

6.1 System configuration

6.1.1 Repeaters

As we mentioned in Section 1.5, the AR6A hot-standby configura-

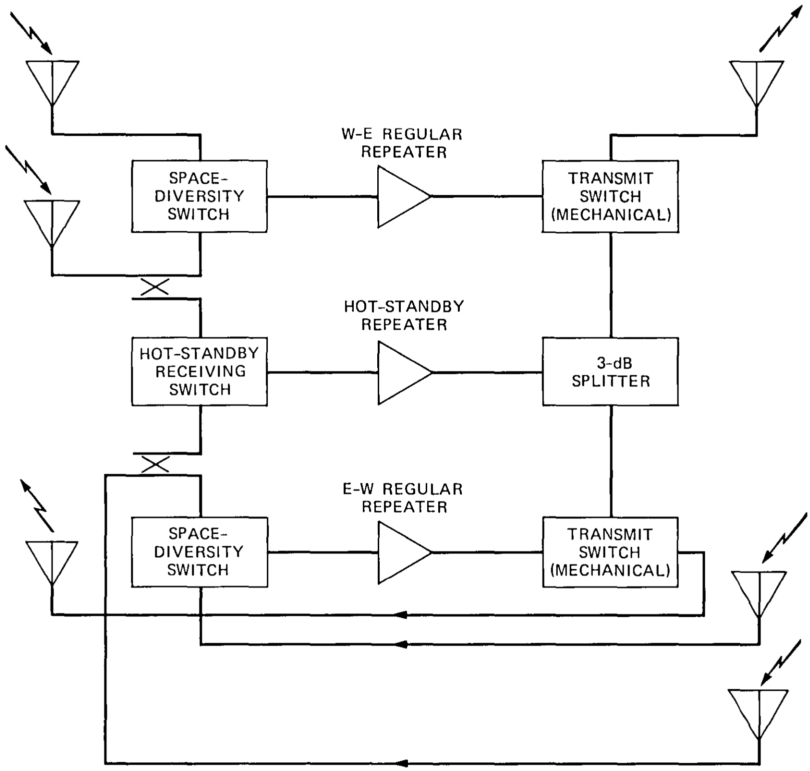

Fig. 19—Hot-standby repeater configuration with space diversity.

tion uses a single hot-standby repeater to protect both directions of transmission. This is illustrated in Fig. 19, which shows a repeater configuration using space diversity.

The primary monitoring function determining whether a TR bay is in satisfactory condition is provided by a broadband power monitor at the transmitter output. Under some trouble conditions associated with high receiver gain, it would be possible for broadband receiver noise to look like a message load, thereby deceiving the power monitor. To avoid this difficulty, high receiver gain operates the pilot resupply switch at the receiver output in the same way as it does in frequency-diversity applications. However, since resupply pilots are not needed for hot standby, the normal pilot input port can be terminated. Thus, when the resupply switch is operated, the transmitter input is automatically terminated, providing the equivalent of a squelch.

Figure 19 shows that in the hot-standby mode the hot-standby receiver does not have space diversity. This eliminates the necessity and complication of transferring the space-diversity control function between regular and hot-standby bays. The simplification is justified on the grounds that the estimated simultaneity of equipment failure and fading is small enough to be neglected.

The microprocessor-based control unit, in addition to monitoring transmitter power output and other status indicators, arranges the sequence of switch operations during the transfer to or from hot standby. In addition, it provides local and remote status information and supervises local and remote commands with predetermined override priorities. Important considerations relating to the reliability of the 1×2 hot-standby configuration are discussed in Section 6.2.

6.1.2 Main stations

Main stations do not lend themselves to a 1×2 hot-standby protection arrangement since in general there is only one TR unit per channel at these locations. As a result, main station protection is provided on a one-for-one basis for both transmitters and receivers. A block diagram of the protection arrangement that is integrated with the 501A Protection Switching System is shown in Fig. 20. In essence, the transmitting portion of the 501A System double feeds two transmitters whose outputs are selected by a transmitting switch supervised by a control unit similar to the one used at hot-standby repeaters. On the receive side, receiver selection is performed by the 501A switch on

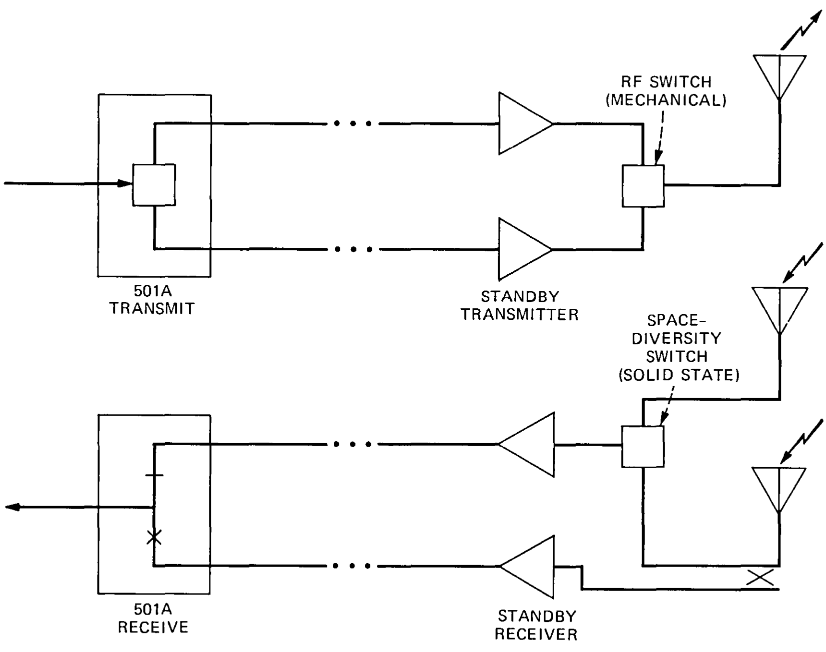

Fig. 20—Hot-standby main station configuration with space diversity.

the basis of noise and pilot monitor information identical to that employed on the 500A System.

Interface levels between the 501A System, TR equipment, and multiplex equipment are identical to those employed in the frequency-diversity 500A configuration. This simplifies the conversion from hot standby to frequency diversity when justified by growth in the route cross section.

6.2 Reliability considerations

The main station 1 × 1 hot-standby configuration follows conventional design practice and will not be discussed in detail. For these systems, the use of high-reliability components and subsystems provide adequate overall equipment reliability.

The outage probability of the repeater station configuration shown in Fig. 19 can be expressed as

$$P_0 = P_B^3 + 3P_B^2(1 - P_B) + 2P_S P_B(1 - P_B)^2 + 3P_S' S_B(1 - P_B)^2 + P_S'(1 - P_B)^3, \quad (8)$$

where

P_0 = two-way equipment outage probability,
P_B = outage probability of a radio bay,
P_S = probability of switching system failures that do not cause immediate service outage, and
P_S' = probability of switching system failures that cause immediate service outage.

An example of a failure included in P_S is a loss of power to the switching system when it is in the normal state with both regular radio bays operable. The system equipment outage objective results in the requirement that $P_0 \leq 3.21 \times 10^{-7}$. Preliminary calculations showed that, for expected equipment failure rates and a 2.25-hour mean repair time,[12] the sum of the first four terms of eq. (8) would be less than the required P_0 with reasonable margin. Preliminary calculations for P'_S, however, showed that the last term would be roughly equal to the requirement and special design features would be needed. The probability P'_S was reduced by requiring two independent circuits to simultaneously cause the electromechanical switch shown in Fig. 21 to connect the hot-standby bay to the antenna. Failures in either circuit then cannot cause a loss of service if the opposite direction of transmission is connected to the hot-standby bay input. With this design feature, the equipment outage objective is met for the calculated equipment failure rates.

For hops having space diversity, it was found that in order to meet equipment reliability objectives the space-diversity switch would have

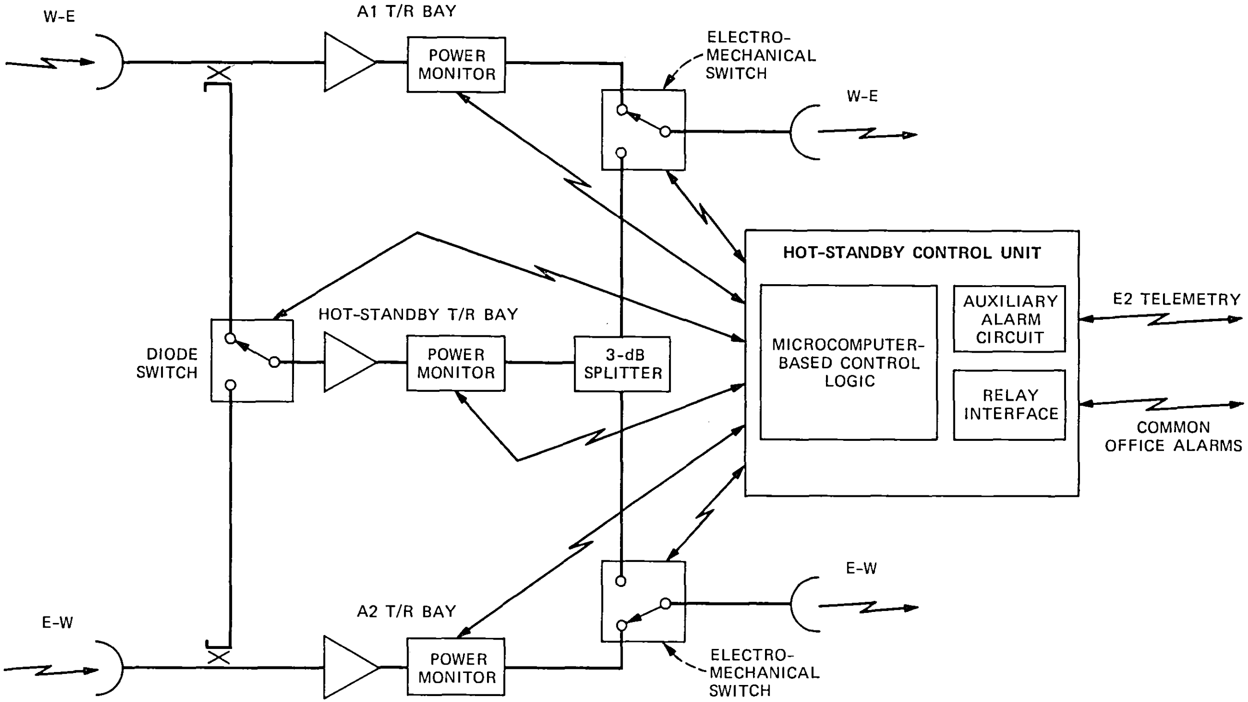

Fig. 21—Block diagram of the 1 × 2 Hot-Standby Protection System.

to be included within the transmission path protected by the standby bay. This leaves the hop without space-diversity protection when the regular radio bay fails but the calculated increase in fading outage is less than 5 percent. The overall reliability of this configuration is better than a design that leaves the switch outside the protected path.

6.3 Switching and control features

A block diagram of the hot-standby switching and control system configuration for repeater stations is shown in Fig. 21. A microcomputer-based control unit is used to control all switching activity and provide equipment checking for alarm purposes. The auxiliary alarm circuit monitors dc power and provides information to the common office alarms. The relay interface conditions the alarm signals for the C1 or E2 telemetry and common office alarm systems.

The hot-standby control unit has three modes of operation: automatic, local, and remote manual. Under normal conditions in the automatic mode, the transmitter electromechanical switches connect the regular (A1 and A2) TR bays to the transmitting antennas. Radio bay failures are detected by the power monitors. A failure of either bay A1 or A2 will cause the system to replace it with the hot-standby bay. For transmitter failures, the loss of signal during switching will normally be less than 15 ms. For receiver failures, the loss will be approximately 50 ms. The standby bay will remain in service until either the regular bay returns to normal or a switch command in the local mode is issued. To revert automatically from the standby bay, the regular bay must be operating correctly for 2 seconds. If another bay should fail when the standby bay is in service, then the appropriate service failure alarm will be set. To prevent unnecessary switching due to failures at a previous repeater, a simultaneous failure of a regular bay and the hot-standby bay is interpreted as an input signal failure and no switching action is taken.

The local mode of operation is provided for maintenance and trouble isolation. The local actions include manual switching of the hot-standby bay to replace either regular bay and a lockout of the standby bay. When the local mode is entered, the pushbutton actions take precedence over switch requests indicated by the power monitors.

If the system is in the automatic mode and the standby bay is idle, the remote manual mode can be enabled by the C1 or E2 telemetry. The remote switching and lockout commands are identical to the local ones. If a service-affecting failure occurs while the system is in the remote mode, a reversion to the automatic mode takes place and the service is protected.

VII. CONCLUSION

The economic attractiveness and efficient spectrum use of the AR6A

System has led to its rapid deployment throughout the United States. The relatively trouble-free introductory experience with this new technology system is a tribute to the many people who contributed to its development. As the authors of this overview article reporting on the work of many colleagues, we wish to acknowledge the contributions of all who participated in the project.

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The AR6A Single-Sideband Microwave Radio System:

Radio-Line Physical Design

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(Manuscript received February 7, 1983)

The physical design of the AR6A radio-line equipment is the product of combining the system operating requirements with the technology available to meet those requirements economically. As part of the description of the radio-line equipment, we discuss the practical considerations for partitioning the circuit block diagrams into realizable units. Modularity is a new feature incorporated into the design of the transmitter and receiver radio bay. We also describe a novel thermal design detail used with discrete transistors in the IF circuits to ensure their reliable operation.

I. INTRODUCTION

AR6A† radio is a new high-capacity, long-haul microwave radio relay system operating in the 6-GHz common carrier band. The use of single-sideband modulation for this high-capacity system was made possible only through the achievement of a high degree of system linearity, which was obtained by designing each circuit within the AR6A System to meet a linearity performance objective. Closely coupled with the system and circuits development of AR6A was the physical design of the radio-line equipment. The Transmitter-Receiver

* Bell Laboratories.

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

(TR)* bay, the TR support bay, and their constituent modules are the principal elements of the AR6A radio-line equipment.

The configuration of this equipment is the culmination of the physical design effort applied to the AR6A System requirements. Physical design is a product creation process that takes a design concept through an iterative development stage and then into manufacture. The driving force behind any new product development, such as AR6A, is the potential market for that product. The AR6A market was identified as the need for a high-capacity, long-haul 6-GHz microwave radio system to overbuild the existing operating companies' long-haul 4-GHz TD-2 and TD-3 radio routes. The principal operating environment for the AR6A radio-line equipment would be within the existing radio station buildings of the TD radio network.

The physical design process began when the AR6A transmitter-receiver was transformed from a systems concept into a functional block diagram. Partitioning the system block diagram into realizable circuit modules was the first step toward the physical design concept. Consideration for the existing buildings and the goals of the operation and maintenance plans was used to determine several fundamental physical design objectives. These objectives were: (1) to assemble a single AR6A transmitter and receiver unit into one radio bay framework, (2) to assemble the common radio-line equipment into one support bay unit, (3) to design for modular field assembly of the TR bay, and (4) to design the radio-line equipment for a nonair-conditioned building environment.

The physical design effort can be divided into two categories: (1) an overall equipment design level, and (2) an individual module design level. The overall equipment design effort is concerned with putting together the many individual circuit modules into one functioning unit, while at the individual design level we concentrated on providing the required circuit functions within the physical bounds determined by the overall equipment design concept. These two categories of design must go on together during development as neither can proceed very far without some knowledge of the other.

The overall radio-line equipment concept was the outcome of an iterative process of configuring each circuit module and then arranging the modules spatially within the framework so that all the physical design objectives were met simultaneously. The particular configuration of each circuit module was influenced by three principal factors: (1) the technology available to implement the required circuit function, (2) the thermal design requirements of the module, and (3) the spatial

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

relationship of the circuit module with respect to the overall equipment concept. The final physical form of each module emerged from the integration of its individual design objectives into the overall equipment concept.

The challenge of this physical design process was to produce AR6A radio-line equipment designs which met both the system performance requirements, and were economically attractive to purchase, to operate, and to maintain.

II. RADIO-LINE EQUIPMENT DESCRIPTION

Figure 1 is a simplified block diagram showing the AR6A transmitter, receiver, and the common radio-line functions. The physical design realizations of this functional diagram are the AR6A TR bay and the AR6A TR support bay.

2.1 Transmitter-receiver bay

The AR6A TR bay, as shown in Fig. 2, is assembled onto a 19-inch panel framework with overall dimensions of 22-3/8 inches wide by 15-1/2 inches deep by 9 feet tall. All the TR bay components are mounted below the 7-foot level in accordance with human factors design considerations. The TR bay is composed of three sections: (1) an upper section consisting of an Intermediate Frequency (IF) shelf assembly and plug-in units, (2) a middle Radio-Frequency (RF) section made up of waveguide-type components, and (3) a lower housing for power supplies and the microwave generator. The predominant color is light gray, which is in harmony with existing radio station equipment. The rest of the color scheme is black, red, and dark blue-gray blended with natural and applied finishes on aluminum, brass, copper, and steel.

The six common alarm and control interface circuits are on printed wiring boards complete with faceplates and are located on the top left level of the IF shelf assembly. A dc-to-dc power converter, which is the common low-voltage power supply for the IF shelf, is located at the upper right-hand side. The majority of the transmitter and receiver IF circuit functions are housed in 12-inch by 12-inch by 2-inch plug-in modules. These IF modules are arranged together for ease of interconnection and operational access into the IF shelf. The transmission path connections are made coaxially across the front of the plug-in modules. A display panel, located immediately below the IF shelf, provides a surface for visual information and dresses the interface between the IF and RF sections.

The receive RF signal path from the indoor waveguide run or from an adjacent AR6A TR bay comes into the channel-separating filter at the 5-1/4 foot level and then down the center of the RF section. With the space-diversity option there are two receive signal paths and two

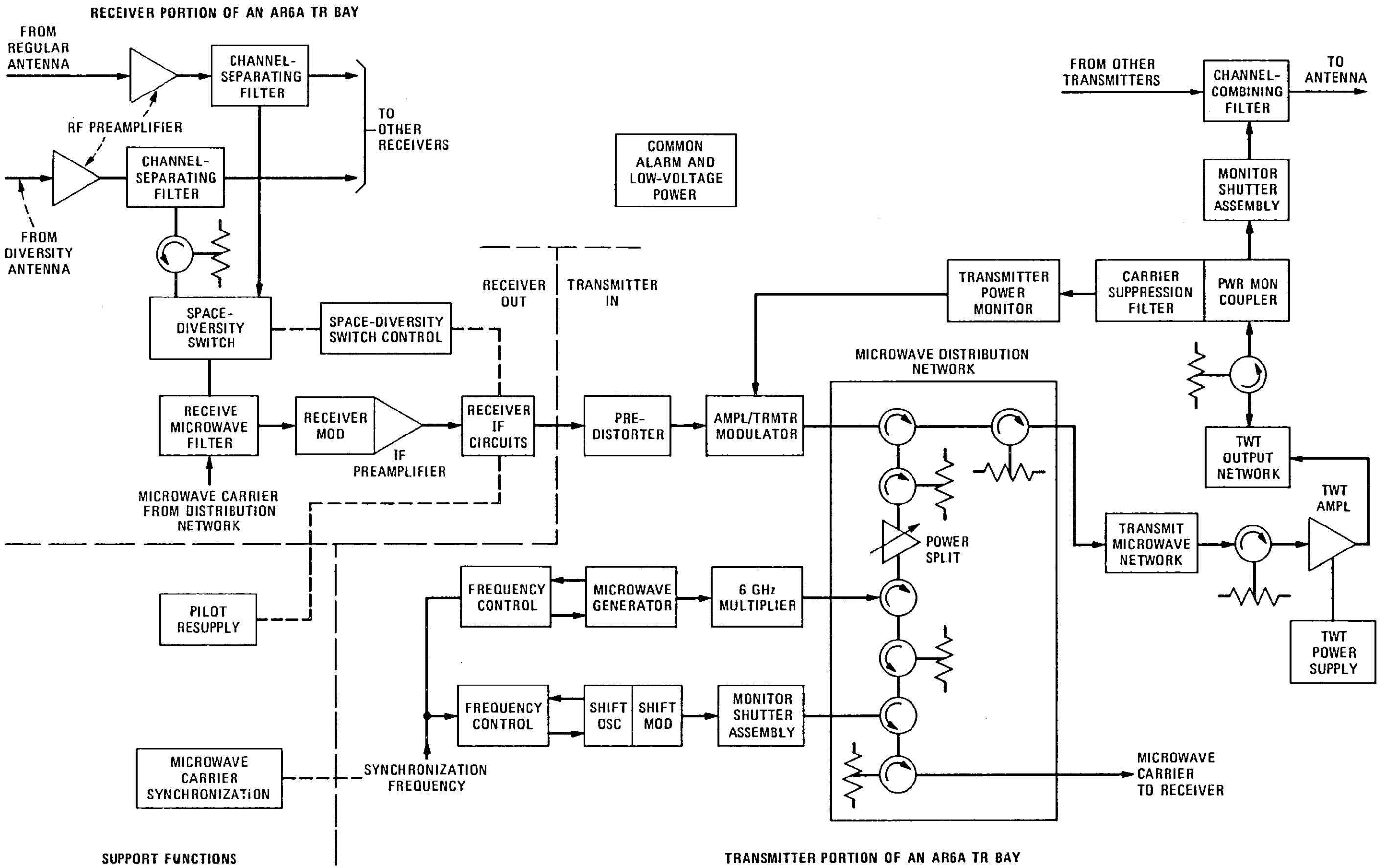

Fig. 1—Block diagram of the AR6A radio line.

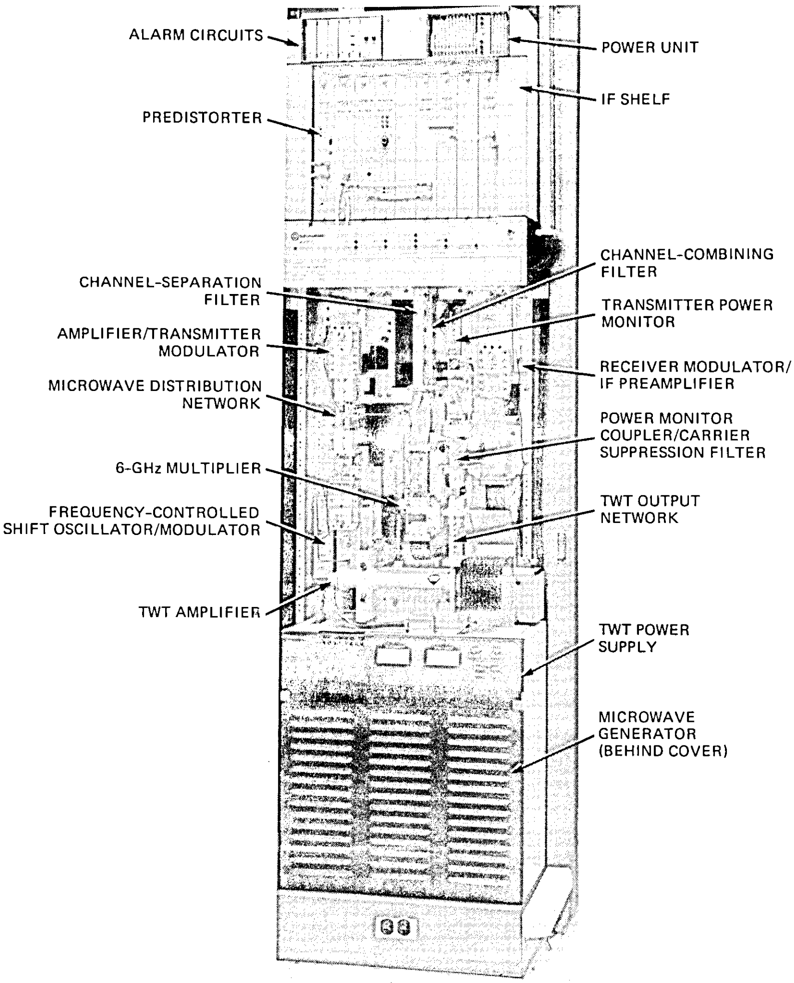

Fig. 2—AR6A transmitter-receiver bay.

channel-separating filters. The upper portions of the RF channel filters are hidden from view behind the removable display panel. The space-diversity switch allows either the RF signal from the main or the diversity antenna to be connected into the receiver. Control of the switch is provided through the switch control unit under the direction of the receiver IF section. The receive signal output from the space-diversity switch is fed through flexible waveguide up to the right, through a waveguide RF filter, and into the receiver modulator. Here the RF signal is down converted to a 59- to 89-MHz IF signal and then amplified in the IF preamplifier unit. This IF signal is cabled up the right to the IF shelf modules where it is passed through each IF module from right to left. The IF modules include filters, amplifiers, and equalizers.

The first element of the transmitter function is the predistorter, which is the leftmost unit on the IF shelf. In a repeater TR bay configuration the output of the receiver is connected directly into the transmitter. The transmit signal, which is still in the IF format, is cabled down the left side of the bay into the IF driver amplifier and transmitter modulator. After the up-conversion process, the resulting signal is transmitted through an RF waveguide filter to the Traveling-Wave-Tube (TWT) amplifier. A transition-to-coaxial cable is used to provide the input to the TWT. The amplified RF signal continues up through a reduced-height waveguide filter and into a waveguide directional coupler. A small sample of the output RF signal is passed through a waveguide filter and then coaxially fed into a power monitor. The RF signal path continues up through a monitor port and into a fixed waveguide-to-coaxial-to-waveguide network. This network makes a compact RF connection to the channel-combining filter. The transmit RF signal path continues out from the combining filter into an adjacent AR6A TR bay or into the indoor waveguide run.

The TWT high-voltage power supply is located in the lower housing directly below the traveling-wave-tube amplifier. The outer shell of the TWT power supply fills the available space for safety and operational considerations. The distinctive shape and color forms a visual transition between the RF section and the lower third of the TR bay. The microwave generator and a –19V dc power regulator unit are mounted behind the cover in the lower housing. The microwave generator is an oven-controlled crystal oscillator with multiplier stages to achieve a 1-GHz output. An external reference frequency is used to stabilize the output frequency. The 1-GHz output from the microwave generator is fed coaxially up to the RF section where a microstrip RF multiplier unit provides the local oscillator signal. This local oscillator signal is supplied via coaxial cable to a distribution network where it is split into two signal paths. One signal path goes directly to the transmitter modulator and the other signal goes to a frequency shifter. The shift oscillator/modulator shifts the transmitter local oscillator signal frequency by 252 MHz to provide the receiver with its local oscillator signal. The shift oscillator is also stabilized by an external reference through a frequency control unit.

2.2 TR support bay

The common functions that are required to operate and maintain the AR6A radio-line transmission path have been configured into a

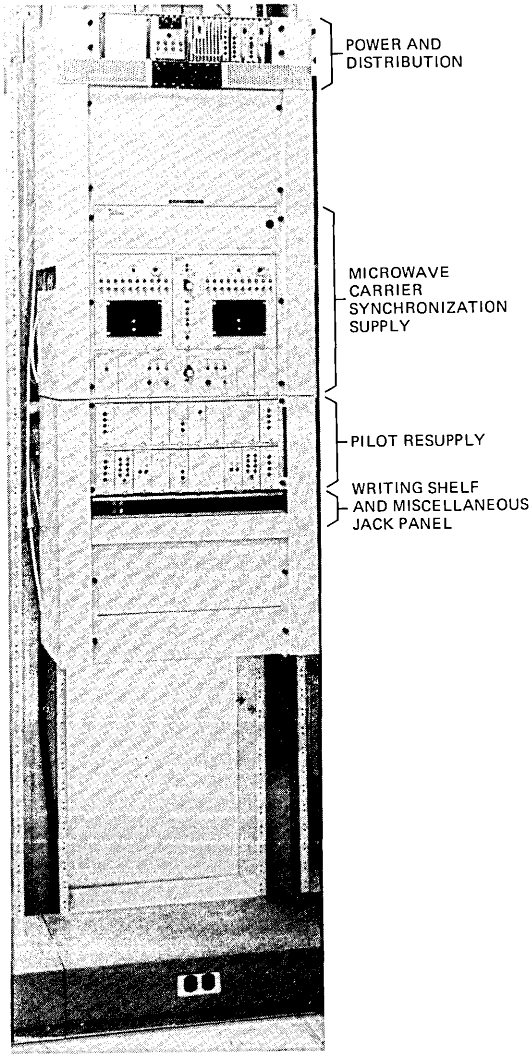

Fig. 3—AR6A transmitter-receiver support bay.

TR support bay unit, which is shown in Fig. 3. These common functions are: (1) the station reference oscillator and its control circuits for synchronization of the local oscillator frequencies within the TR bays, (2) the radio-line pilot resupply oscillators, (3) the additional oscillators and control circuits for out-of-service testing of the TR bays, and (4) the fusing and alarm circuits for the common waveguide RF preamplifiers.

The TR support bay uses the same size framework as does the TR bay so that it can be placed compatibly in the same equipment lineup as the TR bay. Its color scheme is light gray and black in keeping with the TR bay. As with TR bay, all the TR support bay components are mounted below the 7-foot level. The upper shelf assembly provides the mounting for the common TR support bay low-voltage power converters and for the fuse and alarm circuit printed circuit boards associated with the RF preamplifiers. Both the Microwave Carrier Synchronization Supply (MCSS) and the pilot resupply units, configured as shelf assemblies with plug-in modules, are positioned in the upper half of the bay framework for easy access during operation. The panel above the MCSS is reserved for the optional hot standby control unit, while the panels below the jack and access panel provide storage space for the pilot resupply distribution cables.

Additional descriptive information for the TR bay and TR support bay modules is detailed in companion articles in this issue.

III. THE MODULAR APPROACH

An important consideration in any equipment design concept is how the individual parts are put together to form the whole. This section reviews the development of the AR6A radio-line equipment design concept as a modular approach. The AR6A radio-line equipment just described is the result of physical design planning and decision making. The planning and decision process began with the conversion of the system blocks into realizable circuits. Further refinement of the system block diagram produced the evolutional equipment partitioning into the circuit modules.

3.1 Modular design concept

Recall that the physical design objectives included the design of the TR bay for modular field assembly. Station engineering and installation costs may be reduced by initially installing a complete lineup of unequipped, factory-tested, wired bay frameworks in a station, as opposed to engineering each radio-channel addition separately. Channel growth can then be accomplished by installing factory-tested TR modules, as needed, to fill out the radio route. This capability for TR bay modular field assembly provided an added challenge in the development of the radio-line equipment concept.

The initial equipment-concept decision was the selection of the 19-inch panel, duct-type framework for the radio-line equipment. We determined that the floor space available in existing radio station buildings would be sufficient for a full array of frameworks of this size. A smaller framework size would not, therefore, "save" floor space. On the contrary, a more compact framework size might create additional difficulties in engineering for the thermal design, for maintenance access, and for future changes.

A decision on the framework size could not be made without some confidence that its volume would be appropriate for the transmitter and receiver circuit modules. The size and shape of each circuit module was initially approximated based on the proposed technology for each circuit module. Alternate design options were evaluated within the criterion of meeting the overall AR6A System objectives economically. Basic decisions such as these were part of an iterative decision-making process where the impact of a decision on the adjacent levels of the physical design process was fed back to the decision makers.

Various equipment configurations of the TR bay were evaluated as possible design alternatives. The significant factors that influenced the spatial organization of the circuit modules into the final radio-line equipment concept were: (1) the modular assembly objective, (2) the human factors considerations for operation and maintenance, (3) the integration of the overall thermal design concept into the overall equipment concept, (4) the inherent size and shape of some of the circuit modules, and (5) the performance-related objective of minimizing the lengths of the transmission path circuit interconnections.

Modular field assembly was successfully accomplished by structuring the TR bay as the combination of (1) a wired basic bay, (2) a wired IF shelf assembly, and (3) an assemblage of modular RF components. The basic bay is the 19-inch panel, duct-type framework with a minimal amount of factory-installed wiring and a few mechanical piece parts to accept the mounting of the TR bay modules. The installed wiring provides an interface to the radio station dc power and alarm systems and provides the nontransmission path interconnections between the TR bay modules. Some of the mechanical piece parts are factory located with fixtures to assure the proper alignment of the RF components when they are assembled into the basic bay. This bare-bones design approach allows the basic bay to be installed in a radio station with a minimum of additional capital investment.

In the modular design concept, the IF shelf is a wired framework for plug-in modules that has a connectorized interface to the basic bay. Installation of the IF shelf and its plug-ins into the basic bay is a straightforward procedure. In contrast, the RF components did not lend themselves to a traditional modular assembly concept. However, by using the modularity guideline of ease of assembly and by carefully engineering the interface with the basic bay, an innovative modular design was achieved. Each of the RF circuit modules was connectorized, and their mounting arrangements were made easily accessible from the front of the TR bay. The frequency-sensitive waveguide components were assembled together on a supporting framework to minimize the number of waveguide connections that would need to be made in the field. This main RF module is positioned into the basic bay on locating pins for accurate alignment of the channel filters. As an additional dividend from the modular design approach, both the factory assembly of complete TR bays and the field replacement of TR units have been made uncomplicated. The TR support bay has followed the modular design example by using wired shelf assemblies and plug-in modules. There is no advantage in field assembly, however, as the support bay must be installed completely for use with the initial group of TR bays in a station.

3.2 IF circuits subsystem

The IF shelf assembly is a subsystem module within the TR bay as a whole, and, as such, it was treated as a separate entity. A close-up view of the IF shelf assembly is shown in Fig. 4. The subsystem consists of the receiver IF circuits (represented as a single block in Fig. 1) and the closely associated predistortion, alarm and interface, and low-voltage power circuits. These modules are mounted in a two-level shelf assembly that interconnects the modules via backplane wiring. The mechanical design of the shelf assembly is a combination

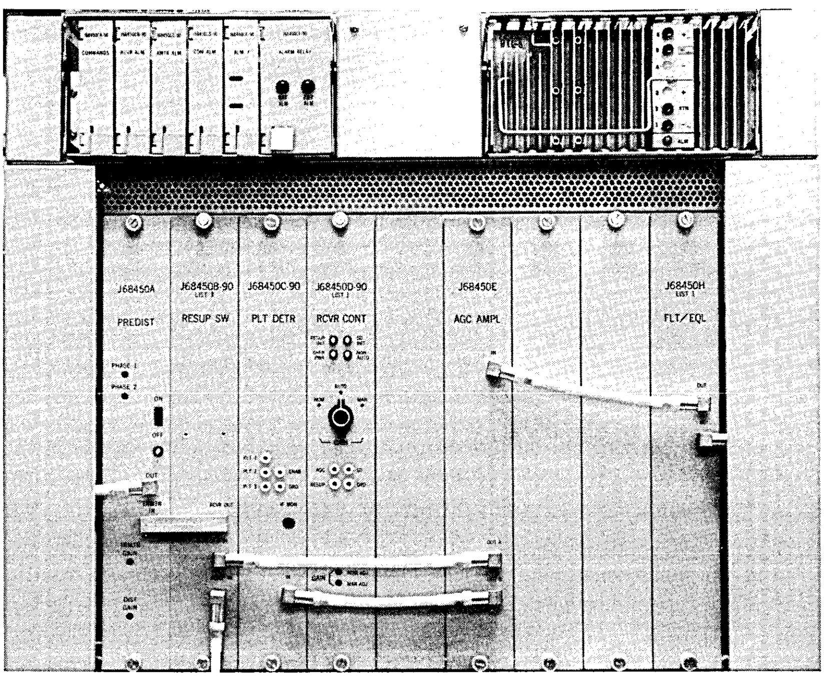

Fig. 4— IF shelf assembly.

of sheet metal fabrication and assembly techniques. Access to the bay framework mounting and the bay cabling is provided by hinged side panels.

A manufacturability cost study showed that for the IF circuits a few large integrated units would be more economical to produce than many smaller units. An electrical and physical partitioning of the receiver IF circuits was achieved based on this study. The block diagram of the receiver IF circuits is shown in Fig. 5. Each of the dashed blocks represents an identifiable circuit function. These circuit functions have been packaged as individual modules: filter/equalizer, dynamic equalizer, Automatic Gain Control (AGC) amplifier, receiver control, pilot detector, and resupply switch. A spare plug-in space is available for future needs. Further description of these modules is found in Ref. 1.

IV. TECHNOLOGY SELECTION

The single most important physical design requirement is that the manufactured product be capable of meeting the AR6A System performance requirements. Therefore, a suitable design technology must be selected for each circuit module that will allow us to meet the individual performance requirements. The choice of design technologies affects the basic partitioning of the circuit functions into physical modules and the total volume needed to realize an equipment design concept. The evaluation of the design technology to implement any circuit function was based on an engineering estimate of: (1) the technical risk factor to achieve the performance requirements, (2) the anticipated cost to manufacture, and (3) the expected reliability.

4.1 Circuit implementation

Circuit linearity is one of the most important of the AR6A performance requirements. The traveling-wave tube was selected for the critical RF amplifier function with the assurance that its inherent nonlinear, third-order characteristic could be controlled and then compensated by predistortion. The performance requirements for the three modulators (receive, shift, and transmit) could be met using modified versions of waveguide structure designs previously used in TH-3 radio.[2] Existing waveguide-based designs were selected as the technology for most of the remaining RF circuit functions. These selections were based on the technical risk confidence factor, performance, and on economic compatibility with the RF section equipment concept. For example, waveguide filters have less loss than dielectric resonator-type filters.

Key elements in the IF circuitry were developed in printed circuit board technology using discrete components. Computer-aided design

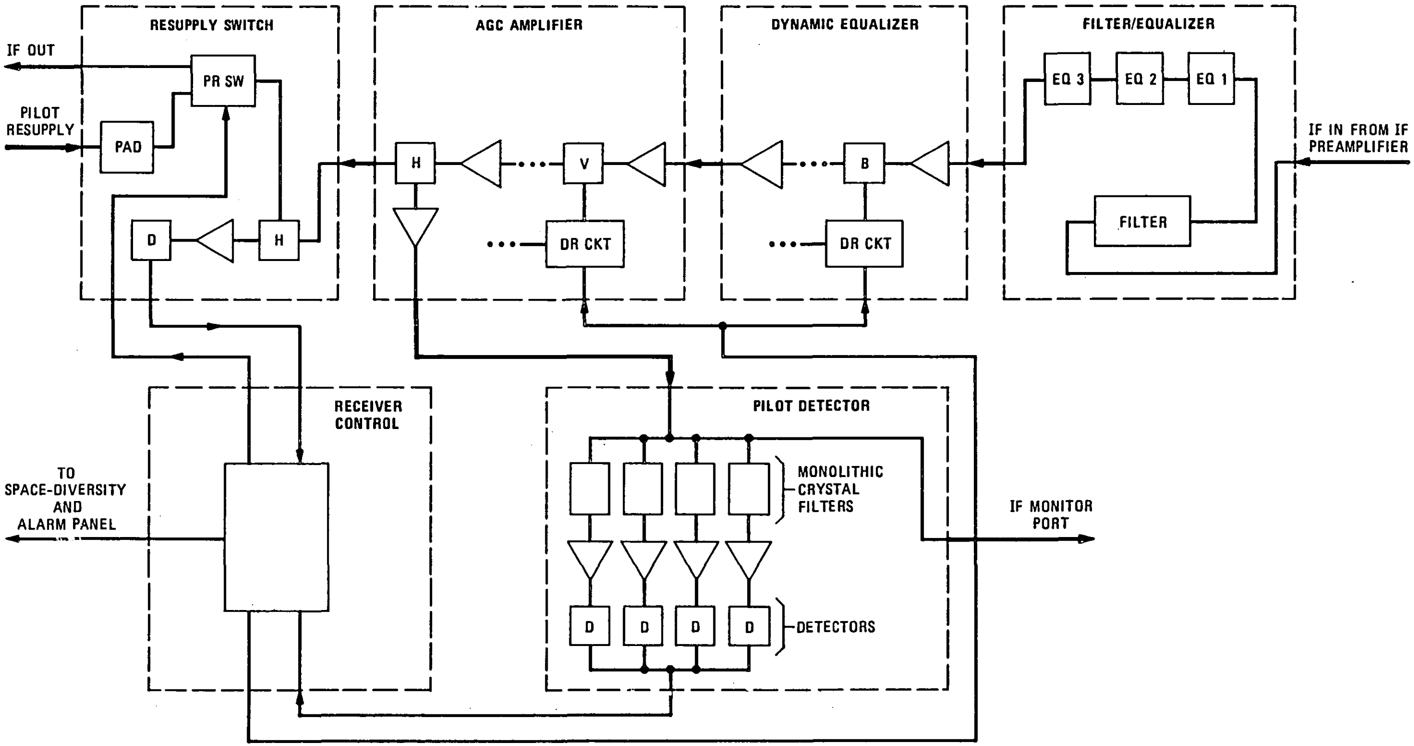

Fig. 5—Receiver IF circuits.

facilities were used to document the critical layout geometries which are characteristic of IF circuits. The performance requirements in the IF circuits were met with amplifier designs based on a hybrid-transformer feedback-transistor stage biased for linear operation. Magnetic components and crystals needed in the IF frequency range were most compatible with the use of printed circuit board technology.

4.2 Mechanical implementation

A technology selection of a different kind was applied to the design of the mechanical piece parts that make up the housings and structures of the AR6A radio-line equipment. The cost of these parts and the labor to assemble them together make up a significant fraction of the total equipment cost. The economics of providing and assembling the mechanical parts is a function of the number of units to be produced. Typically for the radio-line product lines the number of units is not very large, and we use existing manufacturing facilities and tooling wherever possible. Aluminum die castings are used for the microwave assemblies to achieve the required close tolerance forms at relatively low cost due to the reduced need for secondary machining. Aluminum extrusions were utilized as linearly shaped raw material for fabrication into the IF circuit housings. These extrusions can be combined to form electrically isolated compartments within the same plug-in housing. Dip brazing techniques were used to seal those compartment sections that were particularly sensitive to Electromagnetic Interference/Radio Frequency Interference (EMI/RFI) leakage. Close tolerance stamping and forming of sheet metal provides an economical method of creating structural members and parts for the shelf assemblies.

V. THERMAL DESIGN

The critical importance of thermal design to the operation of electronic circuits should not be underestimated. The operating temperature of the components in a circuit affects the performance of the circuit as well as its reliability. The primary objective of thermal design is to maintain the individual circuit components within the temperature limits necessary to achieve circuit performance and reliability objectives. The component temperatures will be determined by: (1) the amount of heat generated by the components, (2) the thermal impedance (°C/watt) between the components and the local ambient, and (3) the local ambient temperature. The thermal environment for the AR6A radio-line equipment is the moderately controlled interior of the radio station. The ambient temperature in a typical, nonair-conditioned, but heated, radio station may range from +4°C (40°F) to +49°C (120°F), depending on the region of the country and the season of the year. It is the upper end of this temperature range that concerns us the most, although some unusual problems may occur at low temperatures.

5.1 Thermal management

Thermal management is a design coordination method that economically achieves the thermal design objectives of the individual circuit modules within the design intent of the overall equipment concept. The thermal aspects of the AR6A radio-line equipment concept were a result of design considerations taken from two perspectives. A total assembly view arranges the circuit modules within the equipment framework to minimize the surface temperature rise of the modules above the station ambient. As part of the initial allocation of physical space within the AR6A TR bay and TR support bay, the thermal design requirements of the individual modules were important decision parameters. The individual module view considers the circuit components as heat sources and devises low thermal impedance paths to the local ambient surrounding each module. In both views physical separation of major heat sources was used to reduce the effects of localized heating on adjacent parts. Potential hot spots were identified by estimating the heat load per volume from the circuit and component information. Natural convection techniques dominate the AR6A radio-line thermal management plan in the total assembly view. In the individual module view, conduction and radiation are combined with natural convection to meet the thermal design objectives. These passive techniques are preferred over aided thermal techniques, such as fans, because of their inherent reliability and no additional energy requirement. Another aspect of the thermal management plan was the design influence directed toward energy efficiency. By reducing the power requirements for the individual circuit designs, we reduced the thermal design heat load as well.

5.2 Thermal design concept

Due in large measure to the thermal management influence, the volume of the AR6A radio-line equipment framework is large enough to rely on natural convection for cooling the component modules. The thermal management plan for the TR bay is to induce air from the station to enter and flow around or through the modules. Conduction paths between the modules and the mounting bracketry aid in effectively increasing the surface area for the ultimate convection heat transfer to the station ambient. The open expanse of the middle RF section allows air from the station to circulate around the waveguide components. Some of this air is drawn up and through the IF shelf, while the rest of the air mixes back into the station ambient. The purposeful up-front position of the TWT provides for an almost direct free-convection transfer of heat to the station ambient. Equipment mounted in the lower housing is cooled by the air that enters through openings in the lower cover, circulates, and then exits up into the rear of the RF section. For the relatively lightly loaded TR support bay, the plan was to design for natural convection from the module face-plates as the principal heat transfer mechanism. Some airflow is induced into and through the shelf assemblies for added heat dissipation.

5.3 Thermal design of modules

The thermal design of each circuit module must be compatible with its other functional requirements. An ideal application of natural convection cooling would be to have the individual circuit components located within unimpeded natural airflow paths. However, because of EMI/RFI requirements, many of the AR6A circuit modules are constructed with completely closed housings. In these modules the power densities are low enough that we can use passive internal heat transfer mechanisms to provide adequate thermal paths. Special attention was paid to the localized hot spots to assure low thermal impedance paths to the outer housing. With a large surface area relative to the thermal load, the heat from these modules can be transferred to the local ambient with a small temperature gradient.

5.4 Special thermal considerations

The largest single heat source in the TR bay is the traveling-wave tube used as the transmitter RF amplifier. Most of the energy in the electron beam is converted into heat at the collector end of the TWT. A critical operating requirement for the TWT is to maintain the temperature of the glass-to-metal seal at the collector below 150°C. This requirement was successfully met by a carefully designed conduction path from the collector to an external-finned heat sink. The thermal impedance between the TWT heat sink and the local ambient was minimized by choosing the optimum fin spacing permitted within the physical space allocated to the TWT.

When a thermal design requirement specifies that a component is to be operated within a narrow temperature range, it may be necessary to control the component's temperature above the ambient temperature by the addition of heat. Several of the frequency-controlling crystal oscillators in the AR6A radio-line equipment are temperature stabilized with thermally insulated oven units. The insulation provides a relatively high thermal impedance path to reduce the heating energy requirements. The control circuit of the oven is set to add no heat at the maximum expected ambient temperature and the full output at the minimum expected ambient temperature.

Two of the closed housing IF modules, the dynamic equalizer and the AGC amplifier, required additional thermal design treatment. Both these modules were provided with perforations on the top and bottom of their frames to allow air to circulate through them. The size of these perforations was selected to use the cutoff waveguide effect to minimize the EMI/RFI leakage. The AGC amplifier dissipates approximately 30 watts of power, mostly from 15 discrete transistors. A vertically finned heat sink has been integrated into the design of the frame for this module. A low thermal impedance path between the discrete transitors and this heat sink is described in the next section.

5.5 Heat-dissipating device

One of the most important circuit designs used in the AR6A System is an IF amplifier stage with very stringent amplitude linearity requirements. To achieve the required linearity, each transistor is biased with a relatively large collector current. Two watts of transister bias power that is converted to heat must be dissipated while the junction temperature is maintained below 125°C. The internal thermal impedance between the junction and the external package of the transistor is 30°C/watt. A patented conduction path package[3] has been devised for this application. Figure 6 shows a cross-section of the heat-dissipating assembly. The purpose of this assembly is not only to provide a low thermal impedance between transistor header and external heat sink but also to provide a means to connect several devices mounted on a printed circuit board to a common heat sink.

The assembly consists of a cylindrical tellurium-copper body, a phenol fiber insulator, a curved stainless steel washer, and a brass ferrule. The body consists of two annular seating surfaces, or "shelves," on the inside surface. The lower shelf provides contact with the transistor, while the upper one provides a stop for the ferrule. Two diametrically opposite pins on the bottom surface of the body engage holes in the printed circuit board to provide rotational locking. The ferrule compresses the spring washer, which exerts a controlled axial force through the insulator to the transistor and, thus, provides a uniform contact pressure between the transistor header and the copper body. A selective indium plating on the inner surface of the body forms a soft interface to improve thermal contact resistance. The insulator provides electrical insulation between the transistor cover and the copper body. The ferrule has a tapped hole for assembly to the common heat sink. It seats on the upper shelf of the body when assembled. An annular groove in the top of the ferrule provides a cylindrical surface for grasping the ferrule during assembly. The outer perimeter of the top of the ferrule is chamfered so that the body may be crimped over for locking. An internal groove in the body simplifies crimping. The

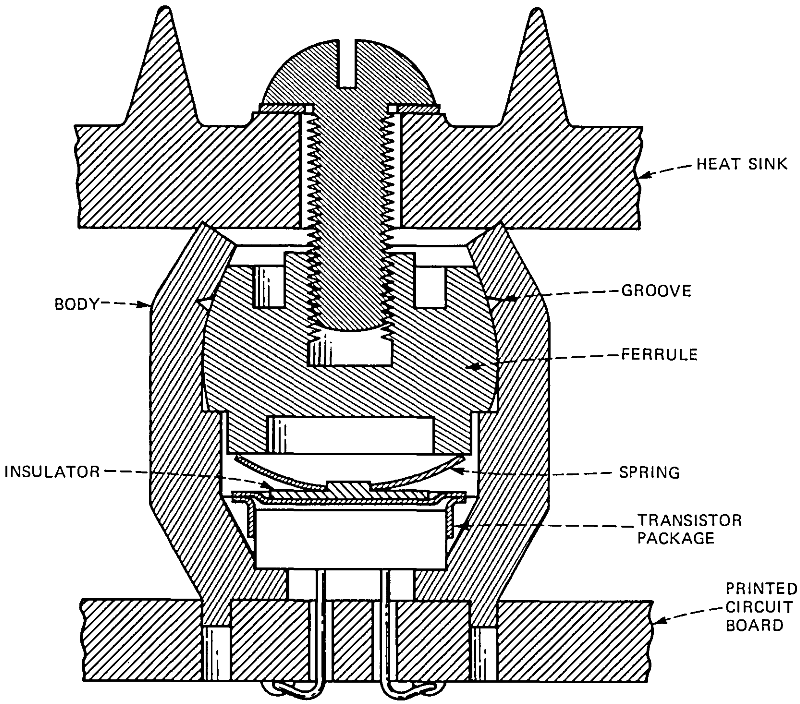

Fig. 6—Heat-dissipating assembly.

tapped hole in the ferrule is "blind" to ensure that the inadvertent use of a long screw does not damage the transistor and to keep metallic debris out of the assembly. When assembled, the ferrule is recessed slightly below the top of the body to provide consistent annular thermal contact between the body and the heat sink.

The material chosen for the body is copper alloy 145. It has high thermal conductivity and good machinability. The free-machining brass of the ferrule is excellent for screw-machine production. Most of the components are designed for screw machining for low-cost production. An assembly fixture presses the ferrule into the body and crimps the body over in a single operation. It also aligns the three transistor leads and two antirotation pins during assembly.

VI. ADDITIONAL DESIGN CONSIDERATIONS

Every product line has a set of standard physical design criteria that all products are expected to meet. Included among these criteria, which have also been applied to the design of the radio-line equipment, are minimum requirements for structural integrity, shock and vibration, and materials selection for nonflammability and corrosion protection.

Beyond these criteria are the special physical design considerations for electromagnetic interference, human factors, and appearance.

6.1 Electromagnetic interference/radio frequency interference

A continued awareness of the potential electromagnetic and radio frequency interference (EMI/RFI) problems was maintained throughout the physical design process. To prevent unwanted signals from propagating into the radio-line circuits, careful attention was placed on the application of interference suppression techniques. Closed conductive housings were used for the sensitive modules and low-pass filters were applied to their dc power and control leads. Grounding locations, and the integrity of those grounds, were particularly reviewed. Wiring and cabling paths were planned to minimize the possible radiated pickup of other signals. These requirements were purposefully blended into the physical fabrication objectives of the modules.

6.2 Human factors

Human factors is a physical design consideration that includes all the interfaces a product may have with people. Of these interfaces, the most important one is safety and personnel protection. Both defined objective safety criteria and subjective value judgments were applied to design details ranging from interlocked high-voltage compartments to the removal of sharp edges and corners from exposed surfaces. Another human factors concern was the operation and maintenance aspect of the radio-line equipment. Panel and faceplate displays should be easily seen and understood by operating personnel, and normal access to the equipment should be unobstructed. On the TR and TR support bays all the important display features have been assigned to favorable eye-zone regions. The AR6A radio-line equipment has been designed for access completely from the front of the bays. Test equipment access points have been brought to front-facing surfaces, and the circuit modules are easily replaced as a part of the modular assembly concept.

6.3 Equipment appearance

Appearance or visual appeal was an important design consideration in the development of the AR6A radio-line equipment. While appearance may not be the final factor in judging any equipment design, first impressions are often lasting impressions. A harmonious design is easier to work with and is given more care by operating personnel. Good design appearance in the radio-line equipment was achieved by incorporating balance into the size, the shape, and the placement of the circuit modules. The same planned attention to detail was applied to the design of the circuit modules to provide functional visual appeal. Color has been used with the radio-line equipment where appropriate to enhance the visual effect. Aesthetically pleasing designs need not be at odds with any other design requirements. Rather, the ultimate objective in physical design is to create simple, but truly elegant, designs.

VII. STATION ARRANGEMENTS

The most economical application of AR6A is to overbuild established TD microwave radio routes. Existing building floor space and facilities, as well as the microwave antennas and towers, can be further utilized at minimal incremental cost. The physical size of the AR6A TR bays was determined with this specific overbuild objective in mind.

7.1 Repeater station—frequency diversity

AR6A uses the same channelization plan as does the TH-3 Microwave Radio System.[4] At a repeater station the two RF signal polarizations in each direction of transmission result in four lineups of four TR bays each. In addition, one TR support bay is required at each station with up to a maximum of 16 TR bays per route. The TR support bay is constrained to be located within 40 cabling feet of any TR bay so that the pilot resupply tone signals may be provided to each TR bay at the same power level.

A typical AR6A repeater station floor plan layout is shown in Fig. 7. Since the radio-line equipment has been designed for front access, the radio bay lineups may be placed against a wall or back to back with another radio bay lineup. The empty bay position opposite the

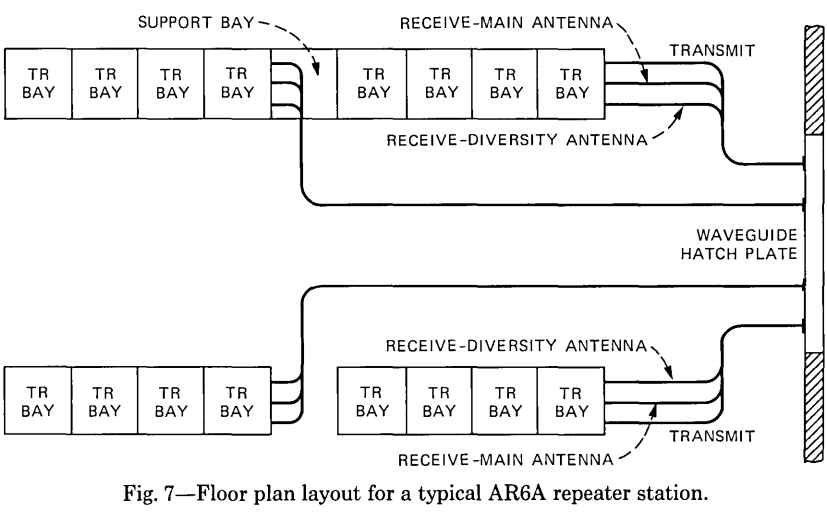

Fig. 7—Floor plan layout for a typical AR6A repeater station.

TR support bay is maintained to provide E-W and W-E channel symmetry across the aisle for operating clarity.

The indoor waveguide arrangements are similar to those used in TH-3. When space diversity is applied there will be three indoor waveguide runs (transmit, main receive, and diversity receive) connected to each bay lineup. To minimize the frequency shift of waveguide filter cavities, dry air is supplied from the station dehydrator using the indoor waveguide for distribution to the TR bays. The indoor waveguide is routed above the bay frameworks and brought down to the 5-1/4 foot level of the channel filters at the end of each TR bay lineup. The common channel RF preamplifiers are mounted in the receive waveguide runs at the top of the first bay in the lineup. Either right-hand or left-hand feed direction is permitted due to the mounting of the channel filters on the TR bay center line. Space has been allocated in the upper section of the TR support bay to allow the indoor waveguide to pass behind the upper side covers. RF connections between the channel filters of adjacent radio bays are made using 4-inch sections of flexible waveguide. Alignment of these connections is made possible by having the radio bay frameworks bolted together through factory-fixtured brackets on the basic bays. Low-pass power filters, which reduce any 60-Hz and harmonic tones from the –24 volts dc power feeds, are located in each bay at the 8-foot level. A photograph showing the AR6A radio-line equipment installed in the Hillsboro, Missouri station is shown in Fig. 8.

7.2 Repeater station—hot standby

When the forecasted channel growth does not permit a frequency-diversity route plan, a hot-standby arrangement is required. Figure 9 shows a hot-standby floor plan for a typical repeater station with one working channel in each direction of transmission. A single standby TR bay protects the two opposite direction regular channels from equipment outage. The receive RF signal paths from the main antenna are split in couplers to provide the inputs from two transmission directions to the standby bay. The microprocessor-based hot-standby control unit, mounted in the TR support bay, controls the selection to an RF switch in the standby bay. The transmit output from the standby bay is split in a coupler above the standby bay and the same transmit signal is sent to another RF switch in each of the regular TR bays. This switch, also under control of the microprocessor controller, determines whether transmit signal from the regular or the standby bay goes out to the antenna. Space diversity may be applied to a hot-standby arrangement but only on the regular TR bays. An additional working channel would share the common receive RF paths but would require a duplication of the other equipment required for the first

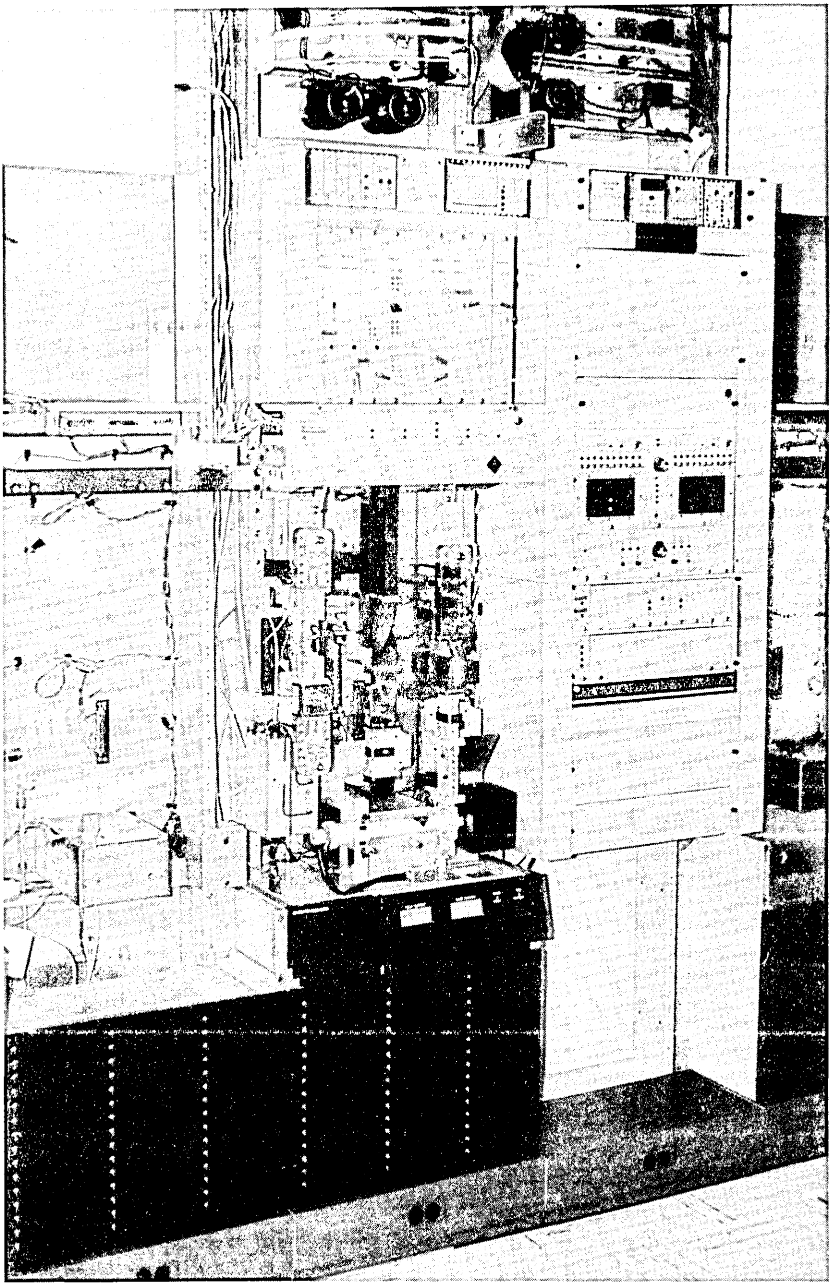

Fig. 8—AR6A radio-line equipment installed at Hillsboro, MO.

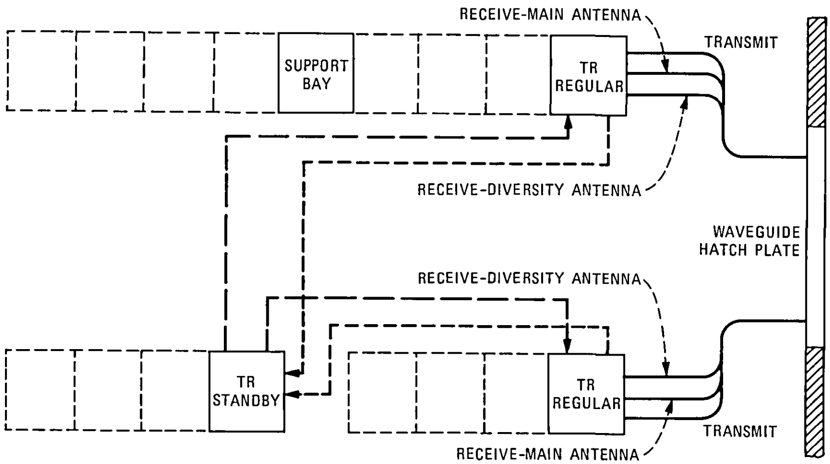

Fig. 9—Hot-standby floor plan layout for a typical AR6A repeater station.

working channel. When there is sufficient channel growth, a conversion to a frequency-diversity arrangement can be accomplished with a minimum of disruption to the working channels. The nonfrequency-sensitive modules from the standby TR bays may be reused in the field assembly of the growth channel TR bays.

VIII. FINISHING TOUCHES

A successful product design is the result of a careful balance of the many system design objectives. Maintaining that balance throughout the physical design process called for increased emphasis on design coordination between the many engineers involved in this major project. Design coordination has become an area of increasing importance in product developments to evaluate design alternatives, to reduce the technical risks, and to keep the project on schedule. Throughout the design cycle of AR6A many formal and informal design review sessions were held to monitor design status and to maintain continuity with the overall design concept. For convenience in keeping all the design engineers informed, the documentation of the equipment design concept was maintained in the format of size-and-feature outline control drawings.

The physical design of the AR6A radio-line equipment will continue to change as technologies improve and new system concepts and circuits are introduced. Improving the product has always been part of the physical design process. In the design of this equipment we have tried to anticipate these future changes so that the physical design modifications may be accomplished gracefully.

It would be unusual in our modern world of design if a complex product, such as AR6A, were created by a few individuals. In addition to the many individuals who were involved in the project, we would like to acknowledge in particular the contributions made by R. E. Caron, G. B. Gregoire, T. Kuliopulos, and L. F. Travis.

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The AR6A Single-Sideband Microwave Radio System:

Radio Transmitter-Receiver Units

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This paper describes the essential features and the performance of the AR6A microwave transmitter and receiver. Subsystems that are major contributors to repeater noise are explained in detail. Design objectives for each of these units are given and compared to typical results obtained from measurements. Transmitter linearity was improved substantially through the use of predistortion. The thermal noise contribution of the receiver was reduced by use of a new low-noise microwave preamplifier.

I. INTRODUCTION

The two basic portions of the AR6A† radio bay are the transmitter and receiver units. These units have a natural interface at the Intermediate Frequency (IF)‡ and are described in the next two sections.

In general, the receiver establishes the required noise figure, and provides the required frequency selectivity, automatic gain control,

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* Bell Laboratories.

† 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.*

and both dynamic and static equalization. The transmitter supplies microwave gain with low intermodulation distortion at the required output power.

The alarm and control system spans the entire bay. This system monitors the operation of the radio bay and provides remote switching capability to exercise important repeater functions to facilitate maintenance testing.

II. THE AR6A RECEIVER

2.1 General

Figure 1 is a block diagram of the AR6A receiver equipped for the space-diversity application. The multichannel Radio Frequency (RF) signals from both the main antenna and diversity antenna (when implemented) are amplified by the optional waveguide microwave preamplifier in each waveguide run. The desired channel is then selected by channel-separation filters. The outputs of the separation filters are input to the space-diversity switch at a nominal full-load power level of −28.6 dBm. The output of the diversity switch is combined with the receive microwave carrier (local oscillator) signal in the receive microwave filter and applied to the Schottky diode single-ended down converter. Integral to the down converter is an IF preamplifier that boosts the signal level to about −4 dBm. In the case when the receiver is not equipped for space diversity, the waveguide from the diversity antenna and the associated channel-separation filter are omitted and the space-diversity switch is replaced by an isolator and a shutter monitor assembly.

The remaining elements of the receiver are located in the IF shelf. The filter portion of the filter/equalizer[1] provides the basic selectivity for the repeater, while the equalizer is used to compensate for static residual amplitude shape due to the various repeater components. At a repeater station the output of the filter/equalizer is applied to the Automatic Gain Control (AGC) amplifier, which provides flat gain with a dynamic range of 61 dB to compensate for fading. Correction is made for both up fades and down fades. The AGC is controlled by the receiver control unit, which maintains the average of the three detected radio pilot voltages constant. The radio pilots are fed to the pilot detector unit through an auxiliary output port of the AGC amplifier. The pilot detector unit filters and detects each of the radio pilots. The output level of each of the detectors is provided as the inputs to the receiver control unit. An IF monitoring point for the radio channel is accessible at the front panel of the pilot detector unit.

At main stations, a dynamic equalizer is used in conjunction with the AGC amplifier to provide up to 16 dB of linear slope or 8-dB

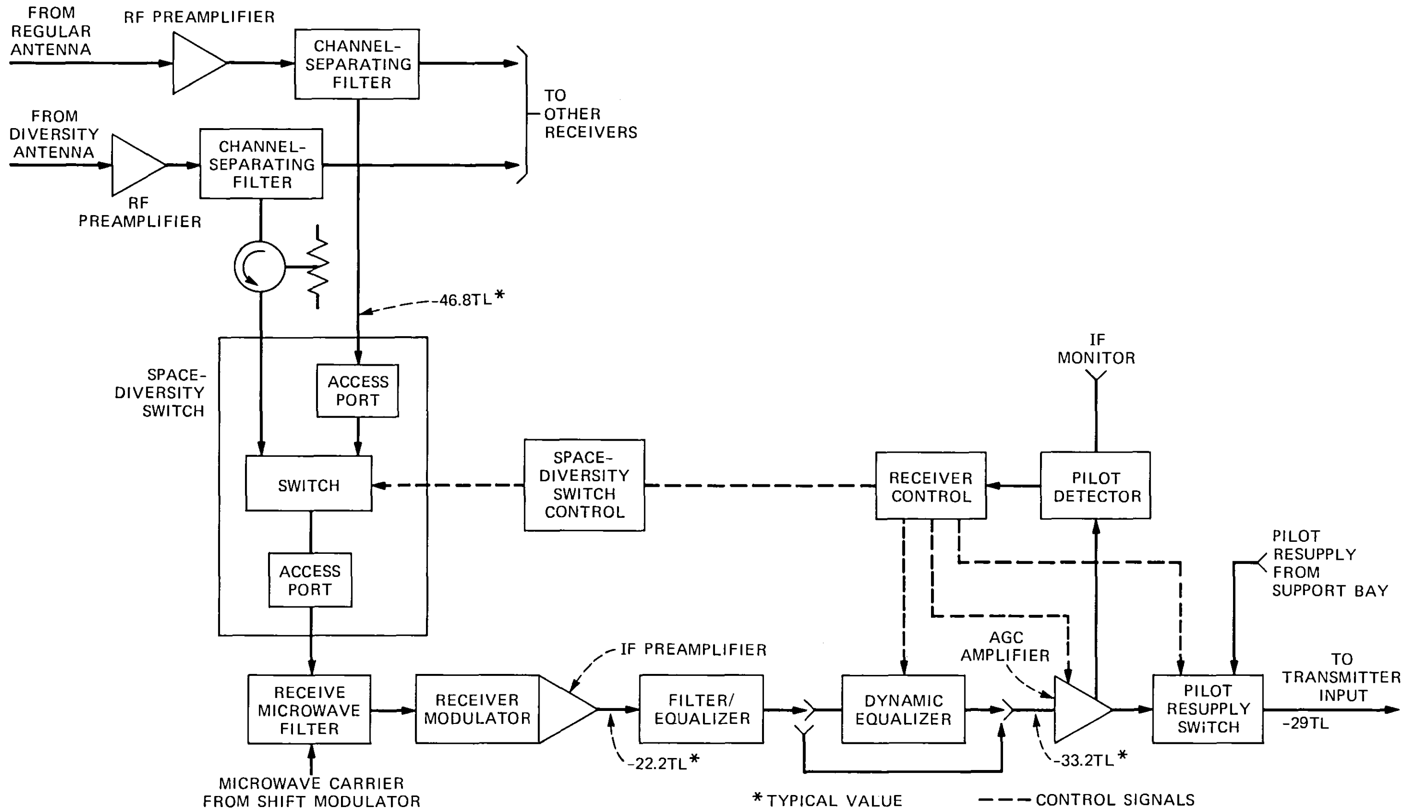

Fig. 1—Receiver portion of an AR6A TR bay.

parabolic amplitude shape to compensate for dispersive fades. The operation of the dynamic equalizer is also governed by the receiver control unit by monitoring the relative differences in level of the three radio pilots.

The output of the AGC amplifier is passed through the pilot resupply switch to the receiver output. Signal loss or signal overpower will cause the receiver control unit to actuate the pilot resupply switch. This action terminates the AGC amplifier output and connects another set of radio pilots from the AR6A support bay to the receiver output. The resupply pilots prevent following repeaters from going to full gain on loss of the channel load.

The receiver control unit also signals the switch request to the space-diversity switch control unit. This request is given when any radio has faded 36 dB or more below the normal received level on a nominal 27.1-mile hop length. This information is derived from the pilot detector signals and the control voltage outputs.

The equalization strategy and operation of the AGC amplifier, dynamic equalizer, receiver control unit, and pilot detector unit are described elsewhere.[2] Further details of the space-diversity switch and control circuit, the receiver modulator/IF preamplifier, pilot resupply switch, and the microwave preamplifier are presented in the following sections.

2.2 Circuit description

2.2.1 Low-noise microwave preamplifier

An optional low-noise microwave preamplifier (656A) is used in the common waveguide run feeding TH-3, as well as AR6A, receivers on a given polarization. The signal on a TH-3 channel is 16 dB higher in power than the full-load AR6A signal; therefore, the amplifier must exhibit excellent linearity to prevent intermodulation "crosstalk" between radio channels. The 656A amplifier intermodulation distortion performance is good enough to allow this common waveguide amplifier to be used with a mix of TH-3 and AR6A on a given route.

The single-stage amplifier using a gallium arsenide (GaAs) field effect transistor (FET) (103B) was designed to operate over the band of 5.925 to 6.425 GHz. The average gain ranges from 8 to 10 dB, with a ±0.5 dB gain shape allowed over the frequency band. A maximum loss requirement of 10 dB is imposed on the amplifier when the transistor is unpowered. The maximum noise figure is 3 dB with typical values measured around 2.7 dB. Typically, the third-order intermodulation coefficient, $M_{A+B-C} = -45$ dB.

Mechanical construction of this amplifier is very similar to the 4-GHz amplifier previously designed for use in the TD Radio System.[3] The GaAs FET is mounted in a microstrip circuit module. Three circulators in air strip-line (Fig. 2) provide good return loss at the input and output waveguide ports as well as an effective passive bypass circuit in case the amplifier module fails. The amplifier has a return loss requirement of 30 dB minimum at both waveguide ports.

The dc operating point for the GaAs FET is a compromise between noise figure and linearity. A built-in regulator sets the gate voltage so that $I_D = 15$ mA at $V_{DS} = 4.8$ volts. The amplifier operates from a −24 volt supply drawing 60 mA. In case of transistor or power supply failure, a contact closure is provided for an alarm indication. At the same time, a visible green Light-Emitting Diode (LED) will be turned off.

2.2.2 Receiver modulator/IF preamplifier

The modulator mixes the received RF signal with the locally generated microwave carrier (LO), which is at a frequency 74.1 MHz above or below the center frequency of the received signal. The modulator, a direct use of an existing TH-3 FM radio design,[4] consists of a stepped waveguide transformer, a waffle-iron low-pass filter, and a diode-mount waveguide section. The basic arrangement of the circuit is shown in Fig. 3. The IF signal is fed through a coaxial low-pass filter to the IF preamplifier input. To obtain improved third-order intermodulation performance for AR6A applications, an LO level of +10 dBm is used instead of +6 dBm as for TH-3. An M_{A+B-C} of about −21 dB is obtained for this LO drive level and optimized diode dc biasing.

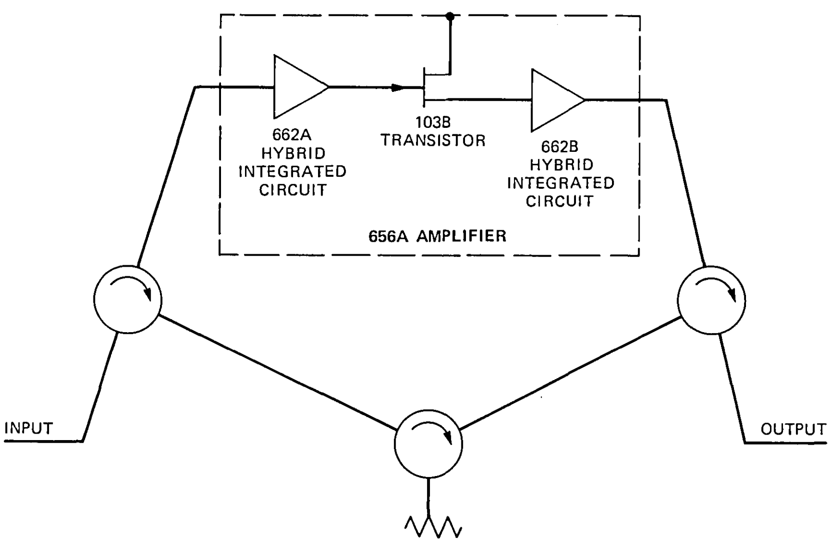

Fig. 2—RF preamplifier.

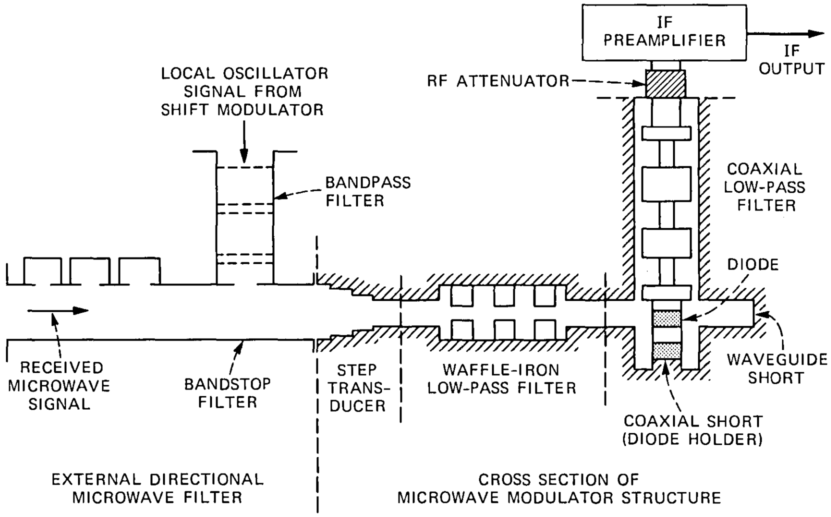

Fig. 3—Basic arrangement of receiver modulator and IF preamplifier assembly.

The 3-stage IF preamplifier has an overall gain of approximately 32 dB. A gain control allows a gain variation of ±3 dB. A slope control is used to achieve flat gain across the band, and two return loss controls are used to provide 30 dB or greater output return loss. The first stage is a low-noise common emitter configuration followed by two hybrid-feedback stages in the output. This configuration provides low-distortion and high-output return loss. The first-stage transistor is designed for low thermal noise performance, but a trade-off was made between best thermal noise and low intermodulation noise. With a dc bias current of about 25 mA and V_{CE} of 3 volts, a gain of 17 dB is realized in this stage with a noise figure of 2.7 dB and an intermodulation coefficient M_{A+B-C} of −45 dB. The following two stages are biased for high-linearity performance. With both transistors biased at an I_c of 120 mA and V_{CE} of 13 volts, the noise figure of the two stages combined is 8 dB with M_{A+B-C} of −80 dB. The complete receiver modulator/preamplifier provides a nominal 28 dB of RF-to-IF gain at a full-load output level of −4 dBm with a noise figure of 8 dB and an overall M_{A+B-C} of −75 dB.

2.2.3 Pilot resupply switch

Figure 4 is a simplified block diagram of the pilot resupply switch circuit. The unit contains a 10-dB directional coupler, a wideband power detector, and an IF switch. The −10 dB port of the coupler is connected through an IF isolation amplifier and power detector. The signal path through the coupler is connected to the receiver-out port

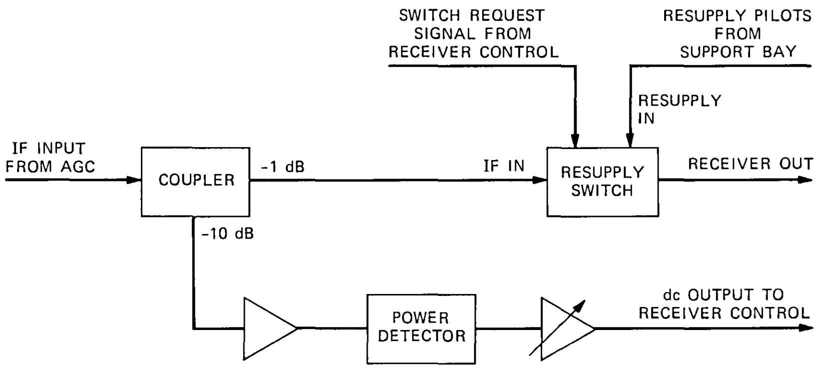

Fig. 4—AR6A resupply switch circuit.

through the normally closed contacts of the switch. The insertion loss through the pilot resupply switch unit is 1 dB. The local set of resupply pilots from the AR6A support bay is applied to the normally open contacts. In this condition, there is 90 dB or greater isolation between the resupply input port and the receiver output port.

The dc output from the power detector is buffered by a variable gain operational amplifier and connected to the receiver control unit. The gain of the amplifier is adjusted to achieve a prescribed output level for a corresponding reference IF power input. This circuit also provides an overpower monitoring feature. If the output exceeds a given threshold for a period of about 80 ms, the receiver control unit will initiate a pilot resupply switch request. The threshold corresponds to an overpower condition of about 12 dB. The receiver control will also initiate a switch request for an underpower condition and in response to a local or remote switch request.

Operation of the resupply switch is controlled by the receiver control unit. When the switch request is made, the IF input from the AGC amplifier is disconnected from the receiver output port and terminated internally in 75 ohms. At the same time the internal termination is disconnected from the local set of resupply pilots, which are then connected through to the receiver output port.

2.2.4 Space-diversity switch and control

The 436A switch, shown in Fig. 5, provides switching between two input ports and a single waveguide output port. One input is WR159 waveguide and is connected to the channel-separation filter that is part of the common waveguide lineup connecting to the main (or preferred) antenna. The other input, an SMA-coaxial port, is connected to the channel-separation filter that connects to the diversity

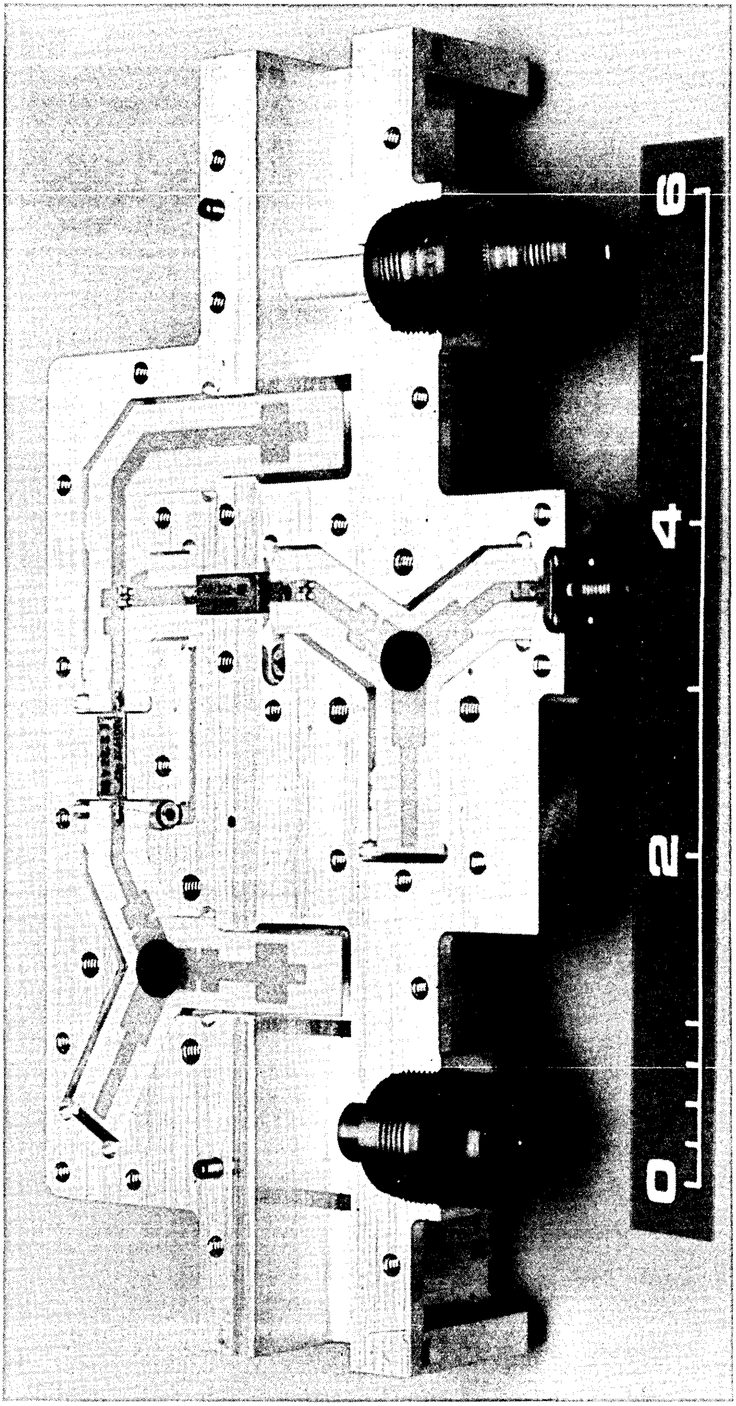

Fig. 5—Space-diversity switch.

antenna. Isolators are built into each port. They provide a matched load to the channel-separation filters when either path of the switch is in its open state. The return loss of the input waveguide port is required to exceed 30 dB; the output waveguide port, 20 dB; and the input coaxial port 25 dB. The insertion loss is required to be less than 1.75 dB. An air leakage path is provided between input and output waveguide ports to provide a path for air when used in a pressurized waveguide system. The switch is designed to fit into the space normally occupied by the 16B shutter monitor and 23A isolator in nonspace-diversity applications. Therefore, the switch is equipped with test access ports and shorting plates to perform the functions of a shutter monitor.

Diode modules in each of the input paths of the switch incorporate three p-i-n diodes. In the preferred signal path the diode module is zero biased to its low-loss (closed) state. In the nonpreferred signal path the diode module is forward biased to its high-loss (open) state. Isolation between input and output ports in the open state exceeds 70 dB.

The space-diversity switch control unit contains the control circuitry for the 436A switch. The circuit can be separated into an analog portion that delivers the driving current function for each p-i-n diode module and a digital portion that includes the fault and alarm circuitry, the reversion circuitry, the main switching circuitry, and the test circuitry.

The switch control panel contains three red LEDs for indicating fault conditions, two green LEDs for indicating which antenna is connected to the receiver, a momentary pushbutton switch for resetting fault conditions or manual antenna selection, and a mode select switch that allows manual or automatic operation.

Although there are no field adjustments, it is necessary to position two miniplugs for selecting various options. One of these user-selectable options allows either antenna to be designated as the preferred antenna and the other enables the automatic reversion option. Other miniplugs are for factory test purposes only.

In operation the main switching circuitry responds to a digital switch request signal generated in the receiver control unit. This switch request will be honored only if the control unit is in the automatic mode and has not switched in the previous 10 seconds. Two driving functions are employed to control the separate diode modules in the space-diversity switch. These driving functions are identical but 180 degrees out of phase and are overlapped to provide a make-before-break type of transition. The maximum rate of switching is limited to one transition every 10 seconds.

If the automatic reversion option was chosen and the receiver has been connected to the nonpreferred antenna for 30 minutes, the reversion circuitry will cause a switch back to the preferred antenna. If a switch request is received during the next 3 minutes, indicating that signal fading is still excessive on the preferred antenna, the control circuit will switch the receiver input back to the nonpreferred antenna and reinitiate the 30-minute timer. If a switch request had not been received during the 3-minute interval, the receiver would remain connected to the preferred antenna.

Any one of six fault conditions will result in a space-diversity switch alarm. Seizure faults are generated whenever either antenna is selected manually or whenever either test function is selected. A reversion fault occurs after 14 consecutive reversion attempts fail. The 15th attempt will cause a 7.5-hour timer to time out thereby inhibiting the operation of the reversion circuitry while allowing the rest of the control circuitry to function normally. A reversion fault is reset by depressing the EXCHANGE ANTENNA/RESET switch. A switch request fault is produced whenever a continual switch request is present for more than 85 minutes. The antenna in use is then seized and further operation of the control is blocked to prevent indefinite switching at a 10-second rate. This fault is reset when a low-to-high logic-level transition is received on the switch request input or the EXCHANGE ANTENNA/RESET switch is depressed.

All control unit input and output levels are Transistor-Transistor Logic (TTL) except the two analog signals driving the space-diversity switch. The ALARM output provides a signal to the alarm and control interface circuit. The output labeled SDS ACTIVITY provides a means of monitoring the space-diversity switch performance.

III. THE AR6A TRANSMITTER

3.1 General

Figure 6 shows the signal-carrying components of the transmitter, together with other ancillary units. The input signal to the transmitter is the 10-mastergroup IF spectrum and includes all pilot tones. This spectrum at a full-load power level of −10.8 dBm comes from either the 500A switch or the output of the preceding receiver and is first applied to the predistorter.[5] The predistorter adds a specific amount of third-order intermodulation distortion with the proper phase relative to the main signal to reduce the intermodulation distortion generated by the remainder of the transmitter. After the signal is predistorted, it enters the amplifier/transmitter modulator, which consists of an IF buffer amplifier, slope equalizer, and a single-ended up converter. The double-sideband RF signal out of the modulator, at a nominal full-load level of −16.7 dBm per sideband, is directed

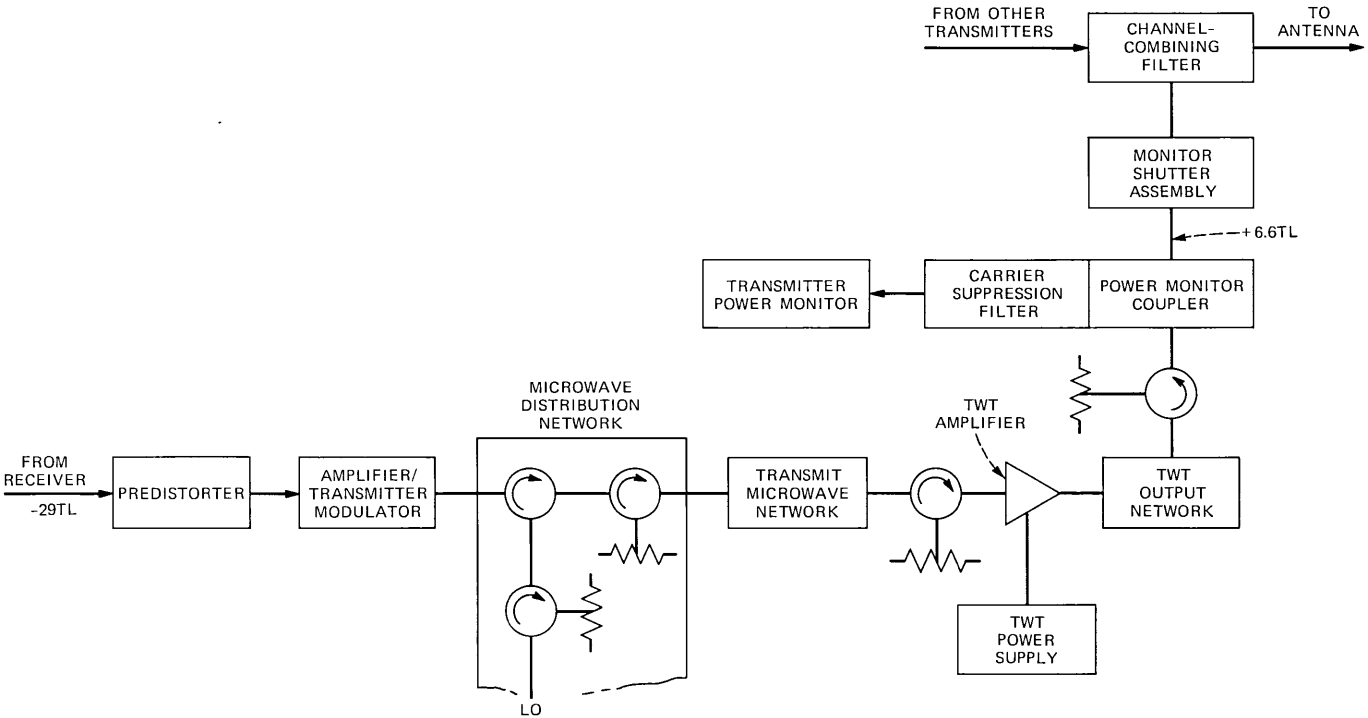

Fig. 6—Transmitter portion of an AR6A TR bay.

through the microwave distribution network to the transmit microwave network (1450 type), which is a bandpass filter with a 1-section delay equalizer. The bandpass filter selects the appropriate sideband and attenuates the carrier and unwanted sideband. More details are given in the next section of this chapter. From the transmit microwave network the signal passes through an isolator to the input of the TWT amplifier.[6] The traveling-wave tube has a typical gain of 44 dB, with gain plus noise figure ≤69.5 dB, and a third-order intermodulation coefficient $M_{A+B-C} \leq -90.5$ dB. A coaxial-to-waveguide transducer is integrated with the tube mount at the input. At the output of the TWT the signal is stripped of its second harmonic content in the TWT output network, which is the combination of a waffle-iron filter, stepped waveguide transformer, and an isolator. The signal is directed through the power monitor/coupler and shutter monitor assembly where the full-load RF power is approximately +25 dBm. The channel-combining filter combines the signal with the spectra of other radio channels.

3.2 Circuit descriptions

3.2.1 The IF predistorter

Figure 7 is a block diagram of the predistorter used in AR6A. After amplification to a level of approximately +1 dBm, the spectrum enters the phase resolver circuit.[1] The phase resolver splits the input signal into two parts with the phase angle between the two output ports being adjustable. The signal passing through the delay line is the main signal and constitutes the reference. The phase-shifted signal is applied to the other path, which is referred to as the cuber path. The delay line in the main signal path is used to match the delay of the other path to within ±0.1 ns.

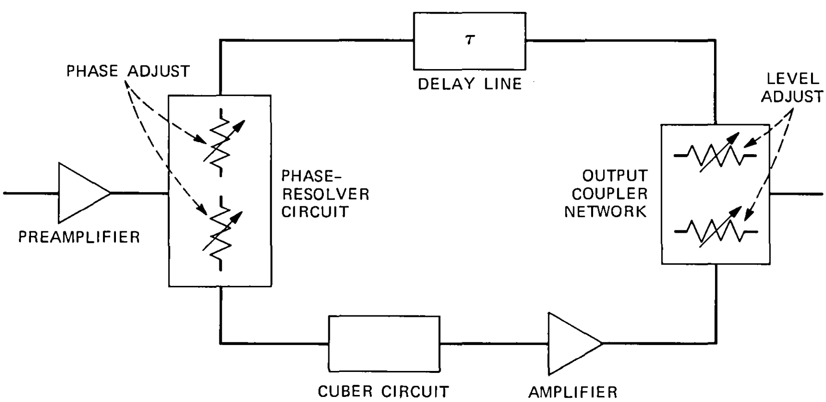

Fig. 7—Block diagram of a predistorter.

The cuber circuit[7] generates the desired complementary distortion products. The cuber circuit consists primarily of two pairs of back diodes connected antiparallel and in opposite branches of a bridge circuit. In this bridge arrangement fundamental frequency signals are canceled by at least 60 dB. The antiparallel pair of diodes generates predominantly odd-order intermodulation products. They are carefully selected for best conversion efficiency and performance that is not level-dependent. Other design considerations and requirements are described in a companion article in this issue.[5] The distortion amplifier brings these complementary distortion products to the desired level and feeds them directly into the output coupler, where the complementary distortions products are combined with the main signal.

The output coupler circuit contains two variable attenuators allowing the independent level adjustment of the main signal and distortion signal prior to combining them. At the output of the predistorter the main path nominal signal level is approximately −14 dBm and the complementary distortion products are at a level of about −71 dBm.

3.2.2 Amplifier/transmitter modulator

The amplifier/transmitter modulator, shown in Fig. 8, is used to translate a 59- to 89-MHz IF signal to the 6-GHz communications band with minimum distortion. The amplifier serves as a buffer between the IF input port and the Schottky diode of the transmitter modulator to minimize variations in the input return loss. Slope and gain compensation of the overall transmission characteristic is accomplished by adjustments located on the amplifier. The amplifier is also equipped with a p-i-n diode attenuator that protects the transmitter output networks by limiting the maximum input power to the transmitter.

The transmitter modulator combines the IF input signal from the amplifier with the microwave carrier. This combining or mixing action results in the creation of a lower sideband (LSB) and a upper sideband (USB) referenced to the microwave carrier.

The amplifier consists of two transistor stages. Ahead of these stages is a transmitter gain adjustment, which sets the transmitter output power by attenuating the IF input signal level. This adjustment is capable of varying the IF signal level over a 20-dB range although, typically, an adjustment less than ±3.9 dB is required.

The first amplifier stage is a common base configuration designed to exhibit very low third-order intermodulation distortion. A slope equalizer network follows the first amplifier stage. When the slope adjustment is rotated clockwise, the network forms a low-pass filter with a negative slope transmission characteristic. When rotated counterclockwise the network becomes a high-pass filter with a positive

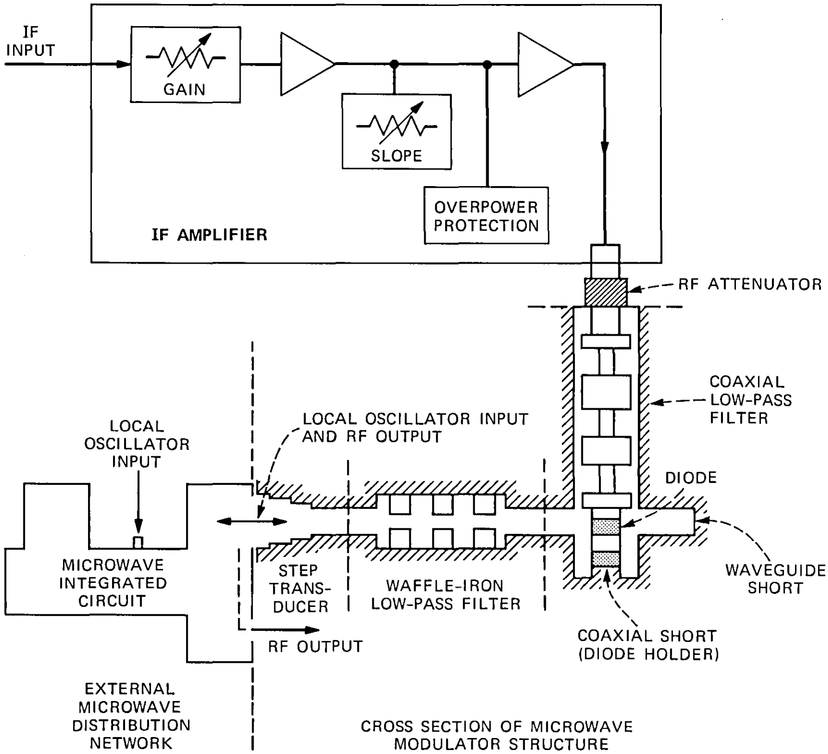

Fig. 8—Basic arrangement of an amplifier/transmitter modulator.

slope transmission characteristic. The slope equalizer network adjustment is capable of varying the IF channel slope over a range greater than ±1 dB although, typically, an adjustment less than ±0.2 dB is required.

The second amplifier stage is a common emitter configuration. It provides approximately 5 dB of additional gain and buffers the transmitter modulator Schottky diode from the slope equalizer network.

The p-i-n diode attenuator network is controlled by a third transistor. Under normal operating conditions, no dc current flows through the p-i-n diode and, therefore, its ac impedance is extremely high. However, should the transmitter output power increase to the point where limiting is required, the transistor will drive dc current through the p-i-n diode and cause its ac impedance to become very low. This loads down the collector of the first-stage transistor thereby attenuating the IF signal by a nominal 10 dB.

The 59- to 89-MHz IF signal from the amplifier and the dc bias for the Schottky diode are fed through an RF attenuator and low-pass coaxial filter to the diode in the transmitter modulator. The coaxial filter suppresses 6-GHz fundamental frequencies and their second and third harmonics while the attenuator provides additional loss at frequencies above 1 GHz to prevent this energy from feeding back to the amplifier. The combined insertion loss of the coaxial filter and attenuator is approximately 3 dB at 800 MHz and 50 dB at 6 GHz.

The microwave carrier frequency is applied to the Schottky diode through a waveguide input port, a stepped waveguide transformer and a waffle-iron filter. The Schottky diode is physically mounted between a coaxial short, which is also the diode holder, and one end of the coaxial low-pass filter. The distance that the coaxial short extends into the diode cavity and the position of the waveguide short are designed to achieve the desired conversion efficiency and eliminate second-harmonic waveguide resonances, which could adversely affect the linearity of the transmitter modulator circuit.

The stepped waveguide transformer provides a smooth transition of the RF transmission characteristic from the reduced height waveguide, used in the design of the waffle-iron filter and diode mount, to full-height WR159 waveguide. The waffle-iron filter prevents the second-harmonic frequencies, which are generated in the transmitter modulator, from appearing at the waveguide port. The waffle-iron filter has an insertion loss of approximately 3 dB at 8.2 GHz and 40 dB at 12 GHz.

The IF signal and the microwave carrier are mixed in the Schottky diode to obtain the RF sideband frequencies. The sideband frequencies appear at the waveguide port and are then filtered to select the desired sideband. The single-tone output power level of the transmitter modulator is nominally −16.7 dBm when a single-tone IF input power level of −13.8 dBm is applied to the amplifier and a microwave carrier power of +16 dBm is applied to the transmitter modulator. A supply voltage of −15 volts dc at approximately 155 mA is required to operate the amplifier/transmitter modulator.

3.2.3 Traveling-wave-tube amplifier

The traveling-wave-tube amplifier supplies the necessary microwave gain. The amplifier has unique linearity requirements: it must provide low third-order intermodulation levels ($M_{A+B-C} < -90.5$ dB) and must have a constant gain and third-order intermodulation coefficient over a wide range of output power (from 23 to 35 dBm).

The amplifier, which is convection cooled, consists of a TWT envelope, focusing mount, and power supply. The input to the amplifier is through an SMA-coaxial connector. The output is reduced height (0.1-inch) WR159 waveguide. Focusing adjustments on the mount minimize helix current.

The power supply provides all voltages required by the TWT. The -24 volt battery, fed through a frame filter to reduce low-frequency (60-Hz harmonics) ripple, is used to operate the high-voltage TWT power supply. Front-panel meters measure helix and collector (or beam) currents. A helix voltage adjustment on the front panel is used to set the TWT to its proper operating point.

IV. MICROWAVE NETWORKS

Various microwave components used in the TR bays were designed some time ago for applications in previous 6-GHz microwave radio systems. Therefore, we are limiting our discussion to new components designed for AR6A.

The 1379-type filter is used as both a channel-separating and channel-combining filter. This filter is mechanically identical to the 1340-type channel-combining filter of the TH-3 System.[8] It is a six-cavity, pseudodirectional filter in WR159 waveguide with a 3-dB bandwidth of 54.15 MHz. To reduce the passband insertion loss slope due to the inductive nature of discontinuities used in the design, the filter is tuned 1 MHz lower than the channel frequency. The insertion loss at the channel frequency should be less than 0.75 dB. In the stopband the loss follows well with the calculated loss of a six-cavity, maximally flat, waveguide bandpass filter.

The receiver modulator filter (1374 type) is a three-cavity directional filter in WR159 waveguide that couples the microwave carrier (LO) and the signal into the receiver modulator. The insertion loss at the LO frequency is no more than 1 dB and the 3-dB bandwidth is 17 MHz. Selectivity at frequencies ±25 MHz away from the center of the bandpass is more than 30 dB. The return loss for both signal and LO ports is required to be at least 30 dB.

The transmitter microwave network (1450 type) is used to block the unwanted sideband at the output of the transmitter modulator. It consists of a seven-cavity, maximally flat, bandpass filter (1373 type) in WR159 waveguide and a microwave delay equalizer.[9] The 1373 filter has a 3-dB bandwidth of 70 MHz. The return loss over the 34-MHz passband is required to be better than 30 dB. The insertion loss at midband is less than 0.65 dB. The delay equalizer adds no more than 0.35 dB to the filter passband loss. The residual delay distortion of the network should be less than ±0.2 ns over a 30-MHz passband.

The TWT output network (1451 type) serves the two functions of transducing the reduced height waveguide at the TWT output port to the full-height WR159 waveguide and of rejecting harmonics from the TWT. It consists of a three-section waffle-iron low-pass filter and a three-section quarter-wavelength step impedance transducer. The network has a 30-dB return loss requirement over the 6-GHz band. The insertion loss at the second-harmonic frequency is about 40 dB.

The 1452 network is a power-monitoring coupler in the transmitter. The main signal path consists of a 26.5-dB cross-guide coupler and a 90-degree step twist. The coupled arm has a two-cavity, band-reject filter to block the LO frequency and a waveguide-to-SMA coaxial transducer in the output port. A 30-dB return loss requirement over the entire frequency band is imposed on the main signal path. The coupled arm requires a 20-dB return loss over a 30-MHz band centered at the channel frequency. About 55-dB rejection is provided at the LO frequency by the band-reject filter.

To reduce multiple reflections in the RF signal path of the radio bay, a long section of bent semirigid coaxial cable in the transmitter was tuned and tested together with its two transducers to WR159 waveguide, as a network (1453 type). It requires 30-dB return loss over the entire frequency band. This network is particularly sensitive because of its connection directly to the 1379-type filter. If a mismatch occurs in the 1453 network, severe ripple may occur in the channel.

Three shutter monitors are also required in the bay: the 16A, used in the transmitter, consists of three sliding shorts and two probes; the 17A shutter-monitor, used in the shift modulator circuit, consists of one sliding short and one probe; the 16B, used in the receiver when the space-diversity switch is not used, is similar to the 16A. The shutter-monitor is a waveguide component that has an access port in the broad wall in which a coaxial probe can be inserted. Slides may be inserted into slots cut into the narrow wall, thus providing a short across the waveguide. The resulting assembly is similar to a coaxial-to-waveguide transducer. With the probe and slides removed, the assembly becomes a simple waveguide section. A minimum return loss of 23 dB can be expected from the coaxial probe, and about 35-dB insertion loss per inserted sliding short can be achieved.

V. ALARM AND CONTROL INTERFACE CIRCUITS

The alarm and control interface circuits consist of six units that provide the alarms, indications, and commands needed for one TR bay. These circuits provide a means for external identification of the particular failure, visual indications of trouble conditions on the TR bay display panel, remote indications of trouble conditions, maintaining visual alarms when audible alarms are silenced or disabled, and converting a command from C- or E-type telemetry equipment to a TTL level for interfacing with the TR bay.

The command unit translates command pulses (generated by a contact closure) from a C- or E-type telemetry system to a latched logic-level change for the remote operation of a function in the TR bay. The command circuits in the bays use a multiplex scheme requiring two command pulses from the telemetry system for the operation of any one function. One command pulse from the telemetry equipment for each bay is used for addressing that bay. This is followed by a second command associated with a particular function. This second command is fed to all other radio bays in both directions of transmission but is accepted only by the addressed bay. It is necessary to readdress the bay for each subsequent operation of a function. The five operations that can be remotely performed by command are as follows: RESUPPLY INITIATE, FLAT EQUALIZER SHAPE, NOMINAL RECEIVER GAIN, GAIN MEASUREMENT REQUEST, and BAY RESET.

Conditions causing alarms and indications are listed in Table I. Inputs to the receiver alarm unit are from the receiver or common equipment only; inputs to the transmitter alarm unit are from the transmitter or common equipment only; and inputs to the common alarm unit are from the equipment shared by both the transmitter and the receiver.

The alarm fuse unit provides two fused sources of -24 volts. One source is used for powering all of the alarm and control interface circuit relays. The other is an additional source of power for the power alarm relays.

The alarm relay unit translates TR bay alarms into contact closures for the operation of external audible, visual, and remote alarms. The

Table I—Conditions causing alarms and indications

Condition	Receiver Alarm	Transmitter Alarm	Remote Indication	Display Panel Indication
Space-diversity switch alarm	X			X
Receiver control alarm	X			
Resupply initiate alarm (delayed)	X			
Equalizer at maximum range (delayed)	X			
Shift oscillator loss of lock	X			X
Radio bay enabled				X
Predistorter off		X		
Low or high transmitter power		X		X
Low dc power	X	X		
Microwave generator low power	X	X		X
Microwave generator loss of lock	X	X		X
Transmitter gain normal			X	X
TWT beam fault			X	
Space diversity initiated			X	
Shift oscillator or microwave generator memory fault			X	X
Resupply initiated by command	X		X	X
Receiver gain nominal by command	X		X	
Equalizer flat shape by command	X		X	

unit also has an alarm cutoff function for silencing the audible alarms. LEDs on the faceplate provide a visual indication for a bay alarm or a power alarm.

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The AR6A Single-Sideband Microwave Radio System:

Terminal Multiplex Equipment

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A combination of new multiplex facilities provides the 6000-message-circuit loading for AR6A radio. It includes a Mastergroup Translator Type-B (MGTB) arrangement that accepts five U600 mastergroups to translate and combine them into a Multimastergroup (MMG) spectrum. A new multimastergroup translator for radio accepts two such MMG spectra from MGTB to produce the intermediate frequency signal and pilots for the AR6A line. A pair of pilots from the Bell System reference frequency transmission unit may be added to the MMG spectrum to be recovered at the receiving end for office synchronization. An office master frequency supply sends synchronization frequencies accurate to within one part in 10^9 for the MGTB and Multimastergroup Translator-Radio translating equipment. Description, design features, and performance characteristics are described in detail below.

I. INTRODUCTION

This section describes the frequency-division multiplex facilities that form and recover the 6000-message-circuit load for the AR6A† Radio System. Starting with the U600 mastergroup (MG)‡ spectrum formed by existing multiplex facilities, terminal equipment recently

* Bell Laboratories.
† 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.*

developed to form the multimastergroup (MMG) spectrum for the L5E Coaxial Cable System is also used to form the MMG spectrum for AR6A. This equipment is the Mastergroup Translator Type B (MGTB).

An assembly of five MGTB mastergroup translators translates for five U600 input signals (0.564 to 3.084 MHz), each having a 2.840-MHz mastergroup pilot. The combined outputs of the five translators form an MMG spectrum to which a 13.920-MHz continuity pilot is added. This MMG spectrum occupies the frequency band of 8.628 to 21.900 MHz.

Two such MMG spectra are used as inputs to the Multimastergroup Translator for Radio (MMGT-R). The transmitting MMGT-R terminal combines each MMG with specific recovery and radio-line pilots and modulates each MMG separately to form the IF spectrum of 59.844 to 88.460 MHz. Figure 1 shows the basic mastergroup, MMG, and IF spectra.

Figure 2 shows the process used in MGTB and MMGT-R terminals to translate and recover the ten U600 basic mastergroups.

An accuracy of one part in 10^9 is required for some carrier and pilot frequencies to generate, administer, and recover the mastergroup, MMG, and IF spectra. This accuracy is achieved by synchronizing the carrier and pilot generators to the 2.048-MHz Bell System frequency standard. For locations where this signal is not available, two reference pilots generated by the Bell System Reference Transmitting Unit (BSRTU) are added to the first MMG of an AR6A channel. The precise 64-kHz frequency difference between these pilot tones of 11.200 and 11.264 MHz is derived from the nationwide Bell System Reference Frequency Distribution System. The Bell System Reference Tone (BSRT) receiver unit retrieves the 64 kHz from these pilots for synchronization of the Office Master Frequency Supply (OMFS), which serves the MMGT-R and MGTB facilities.

II. MGTB TRANSLATOR

2.1 Introduction

The block diagram of the mastergroup translator is shown in Fig. 3. Each translator provides modulation and demodulation, serving both directions of transmission. The same basic design is used for all five translators for AR6A. Only the carrier generator frequency and the bandpass filters in the modulator and demodulator are changed to achieve the desired translated spectrum.

2.2 Improved spectrum utilization

Earlier designs of mastergroup multiplex equipment translate and combine up to six mastergroups to form an MMG spectrum with

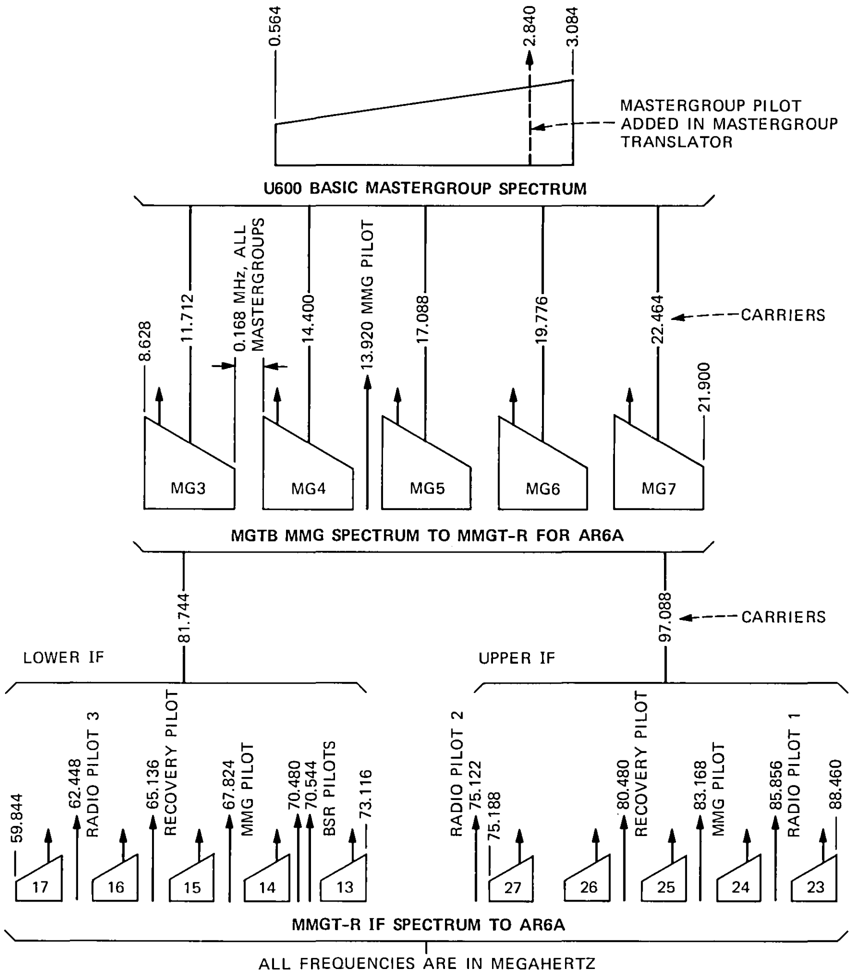

Fig. 1—AR6A terminal equipment spectra.

unequal spacing between the mastergroups. Use of that equipment permitted transmission of 18 mastergroups (10,800 voice circuits) per single coaxial cable or four mastergroups (2400 voice circuits) per FM radio channel.

For the L5E System[1] MGTB equipment translates and combines up to eight mastergroups with a fixed spacing of 168 kHz between mastergroups in the MMG spectrum. By applying three of the MGTB MMG spectra to associated equipment, a total of 22 mastergroups (13,200 voice circuits) are transmitted over a single coaxial cable.

This concept of improved spectrum utilization was carried over into the AR6A development by using five MGTB mastergroups (3 through

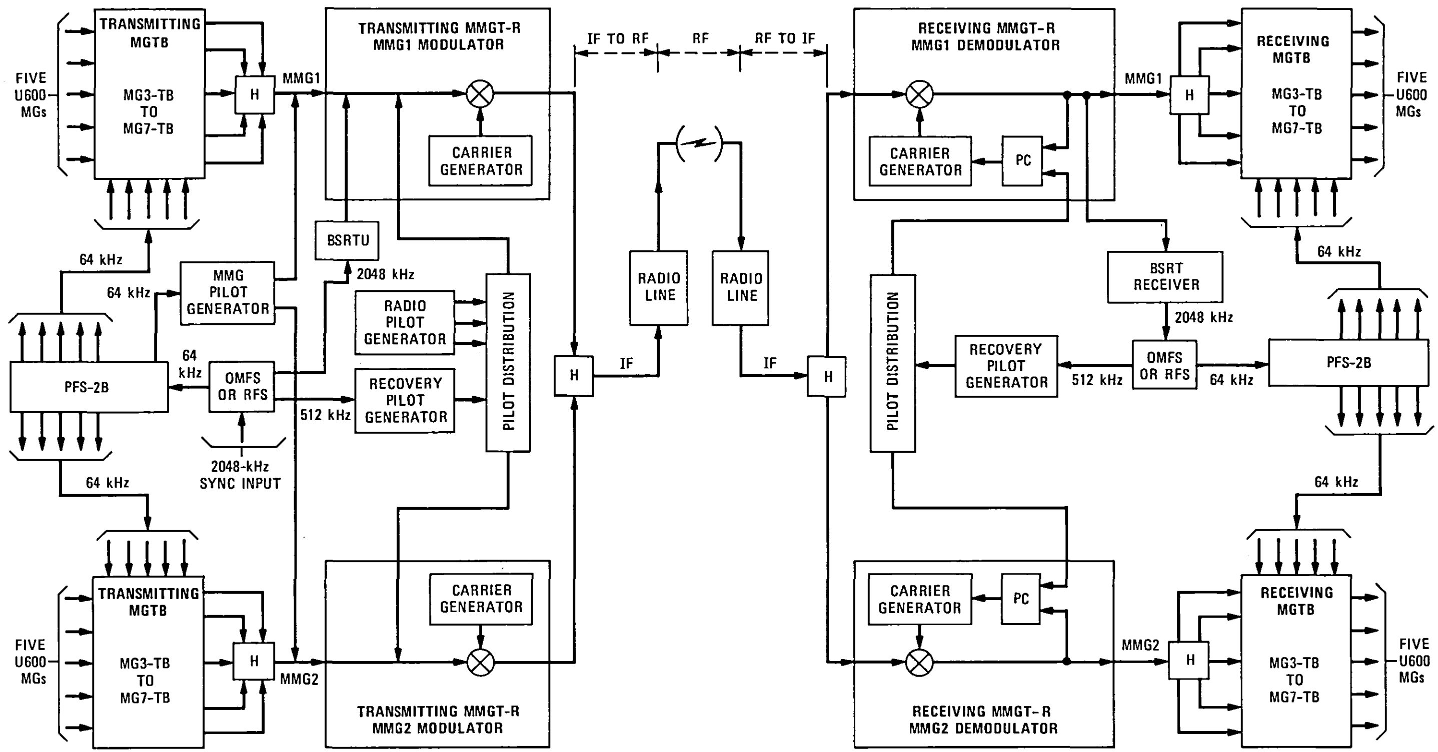

Fig. 2—Block diagram of AR6A terminal equipment.

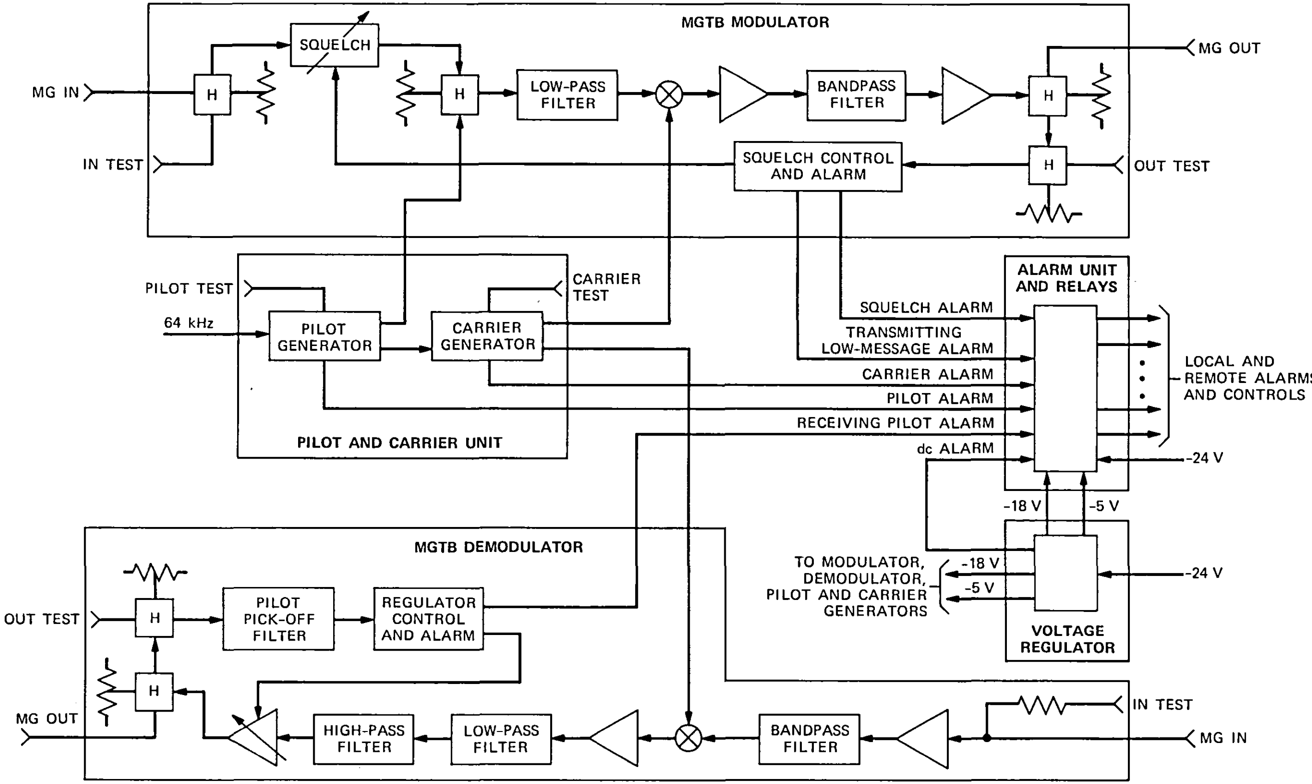

Fig. 3—Block diagram of mastergroup translator shelf.

7) to form the MMG spectrum for input to the MMGT-R. These particular mastergroups were selected to facilitate the realization of filters within the MMGT-R equipment.

2.3 Small size and low initial cost

With MGT designs, the only common equipment in a bay is its battery, fuse, and synchronization distribution panel. Cost of this common equipment is a small portion of the cost of a fully equipped bay. Hence, the incremental growth cost per mastergroup is essentially the cost of the MGT shelf itself. An MGTB bay occupies only half the space needed for earlier equipment designs providing equivalent capability. All circuits for an MGT mastergroup are contained in five small plug-in units per shelf, permitting simplified maintenance or replacement.

2.4 Overall design features

The five plug-in units of an MGTB mastergroup translator are a voltage regulator, alarm unit, pilot and carrier generator, and modulator and demodulator units. These five units contain the circuitry required to translate the U600 mastergroup signal to and from its line spectrum. Test jacks on the plug-in units permit signal monitoring at the inputs and outputs of the modulator and demodulator and at the outputs of the mastergroup pilot and carrier generators.

Transmission and circuit function alarms are displayed locally by indicators on the alarm plug-in and atop the bay. They are also made available to remote surveillance systems. These alarms may be operated in a latched mode, which causes retention of alarm indications until a memory release function is performed.

During a persisting alarm state, the audible alarms are silenced by depressing an Alarm Cutoff (ACO) switch on the alarm unit.

2.5 Circuit design features

2.5.1 Voltage regulator circuit

A −24 volt office battery input drives two series voltage regulators to provide −18 and −5 volt outputs to power the mastergroup translator shelf circuits. The dc drain on the office battery is 1.3 amperes per MGTB shelf.

2.5.2 Pilot and carrier generator circuit

A −20 dBm0, 2.840-MHz mastergroup pilot is used to indicate the relative signal level of the mastergroup. This pilot is added in between supergroups of the mastergroup signal in the transmitting circuit before modulation. Thus, it appears in the line spectrum translated by the carrier frequency. This pilot is used for automatic regulation and alarming. The pilot generator oscillator is phase locked to a stable Primary Frequency Supply (PFS) synchronization source which is locked to an OMFS. Hence, the stability and frequency accuracy of the mastergroup pilot are essentially the same as those of the OMFS supply.

The 64-kHz synchronization signal from the primary frequency supply is divided down to 32 kHz. This 32 kHz is used as a reference signal at one input of a phase comparator. The output of a voltage-controlled crystal oscillator, also divided down to 32 kHz, is fed to the second input of the phase comparator. Phase comparison of these two signals results in a dc control voltage to the oscillator to alter its frequency until the phases of both 32-kHz signals at the comparator inputs are identical. This constitutes the phase-locked loop concept of signal generation wherein the generated signal is stated as being phase locked to a reference signal. A filter at the generator output suppresses harmonics of the pilot frequency.

The MGTB carrier generation is similar to that of the mastergroup pilot generation previously described. The same 64-kHz synchronization signal used for pilot generation is used directly to phase lock the carrier generator to the Bell System reference signal. Two outputs at +10 dBm serve the modulator and demodulator in each MGT shelf. (See Table I for characteristics and performance of MGTB.)

Table I—Characteristics and performance of MGTB

Characteristic	Performance
Type of modulation	SSB-SC-AM
Number of VF channels	600
Basic MG spectrum (U600)	0.564 to 3.084 MHz
MG3B spectrum	8.628 to 11.148 MHz
MG4B spectrum	11.316 to 13.836 MHz
MG5B spectrum	14.004 to 16.524 MHz
MG6B spectrum	16.692 to 19.212 MHz
MG7B spectrum	19.380 to 21.900 MHz
Return loss, all ports across their spectrum	>20 dB
Back-to-back frequency response	flat within ±0.4 dB
Receive pilot regulation range	±5 dB
Loss of pilot regulation	Receiver reverts to normal manual gain setting when input pilot level deviates >−10 dB from nominal.
Noise performance (fully loaded and back to back)	<10 dBrnc0
Phase-jitter performance (back to back) with two independent carrier supplies	<1° peak to peak, Bell weighted
MG carrier phase jitter	<0.7° peak to peak, Bell weighted
MG carrier supply frequency stability	1×10^{-9} with PFS-2B supply and OMFS
dc current drain per MGT shelf	1.3 ampere, maximum at −24 volts dc
dc supply voltage	−20 to −28 volts dc

III. MMGT-R TRANSLATOR

3.1 Introduction

The MMGT-R provides frequency translation between MMG spectra at the MGTB translators and the IF spectrum for AR6A radio. Two MMG signals (MMG1 and MMG2) are individually translated and combined to form the lower and upper portions of the AR6A IF spectrum centered at 74.1 MHz. Three radio-line pilots, required for regulation in the AR6A System, are generated in the MMGT-R transmitter.

Because the signal output of the MMGT-R transmitter is single sideband, suppressed carrier, and amplitude modulated (SSB-SC-AM), minimal carrier energy is transmitted with the sideband information. To enable the receiving translator to track and reconstruct the transmitted message accurately, a substitute form of carrier is added to the message in the transmitting translator. This is a highly stable, 16.608-MHz recovery pilot that serves as an input to a phase-locked loop in the receiving translator to control its demodulator carrier frequency. This corrects frequency offset on message due to repeatered transmission.

MMGT-R terminals are protected on a basis of one protection T and R translator pair for up to 14 regular T and R translator pairs.

3.2 Overall description

3.2.1 Regular terminal

The block diagrams for the regular transmitting and receiving translator portions of the MMGT-R terminal are shown in Figs. 4 and 5, respectively.

In the transmitting direction, MMG1 and MMG2 with 13.920-MHz continuity pilots from MGTB are each translated independently up to adjacent frequency bands of the radio IF spectrum. Their combining produces a composite IF signal of two uncorrelated spectra of 3000 circuits each in the band from 59.8 to 88.5 MHz, centered at 74.1 MHz.

A number of pilots for functions required by MMGT-R and AR6A are generated within the MMGT-R transmitter and are added to the MMG signal at a −10 dBm0 level prior to translation. These include three radio-line pilots at 11.232 and 21.966 MHz in MMG2, and 19.296 MHz in MMG1. Their translated positions in the IF spectrum are shown in Fig. 1. They control automatic gain and dynamic amplitude equalizers in the radio receivers. They also serve as continuity pilots for the 500A Protection Switching System.

A 16.608-MHz recovery pilot is added to both MMG1 and MMG2 spectra prior to translation in the MMGT-R transmitters. This pilot is used in place of the suppressed carrier for demodulator carrier

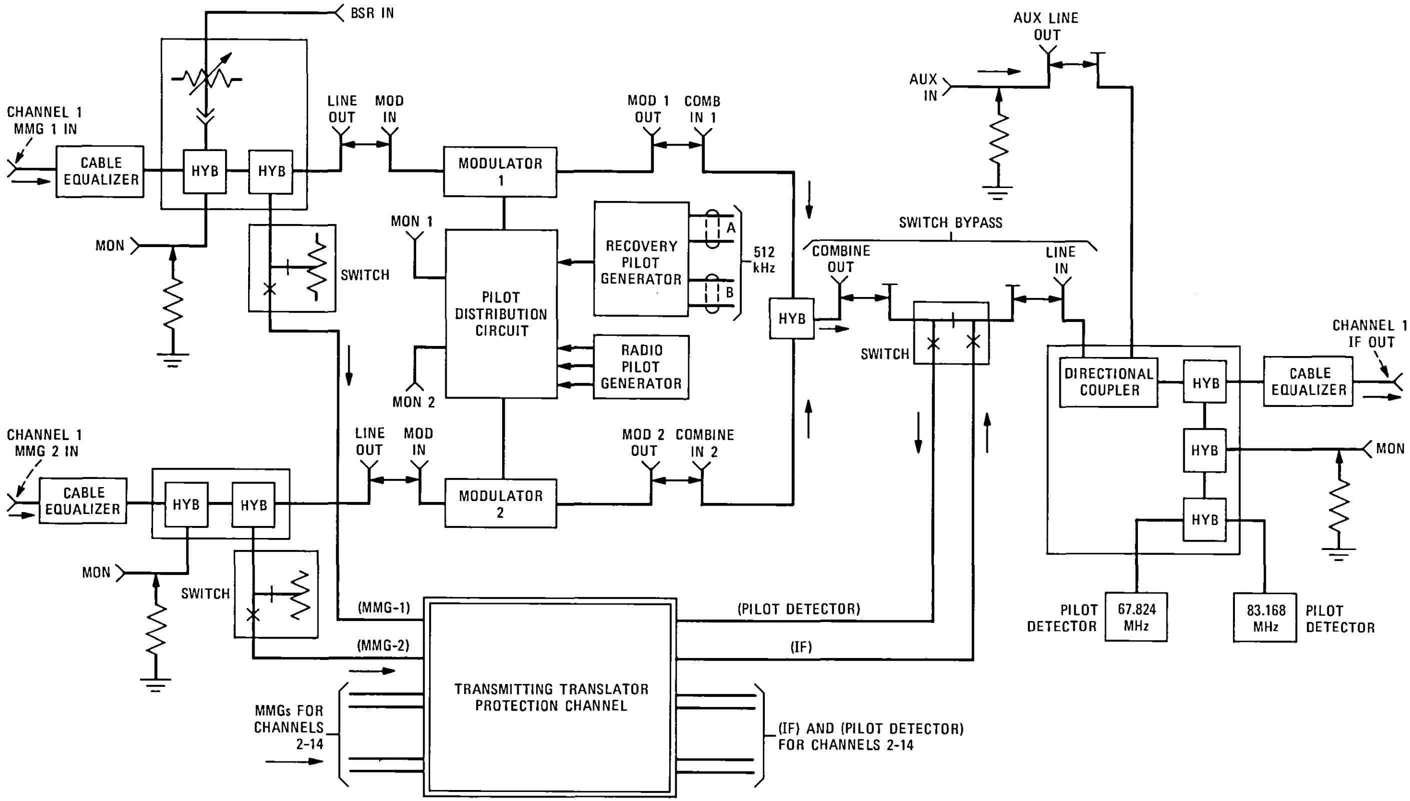

Fig. 4—Simplified block diagram of transmitting MMGT-R translator.

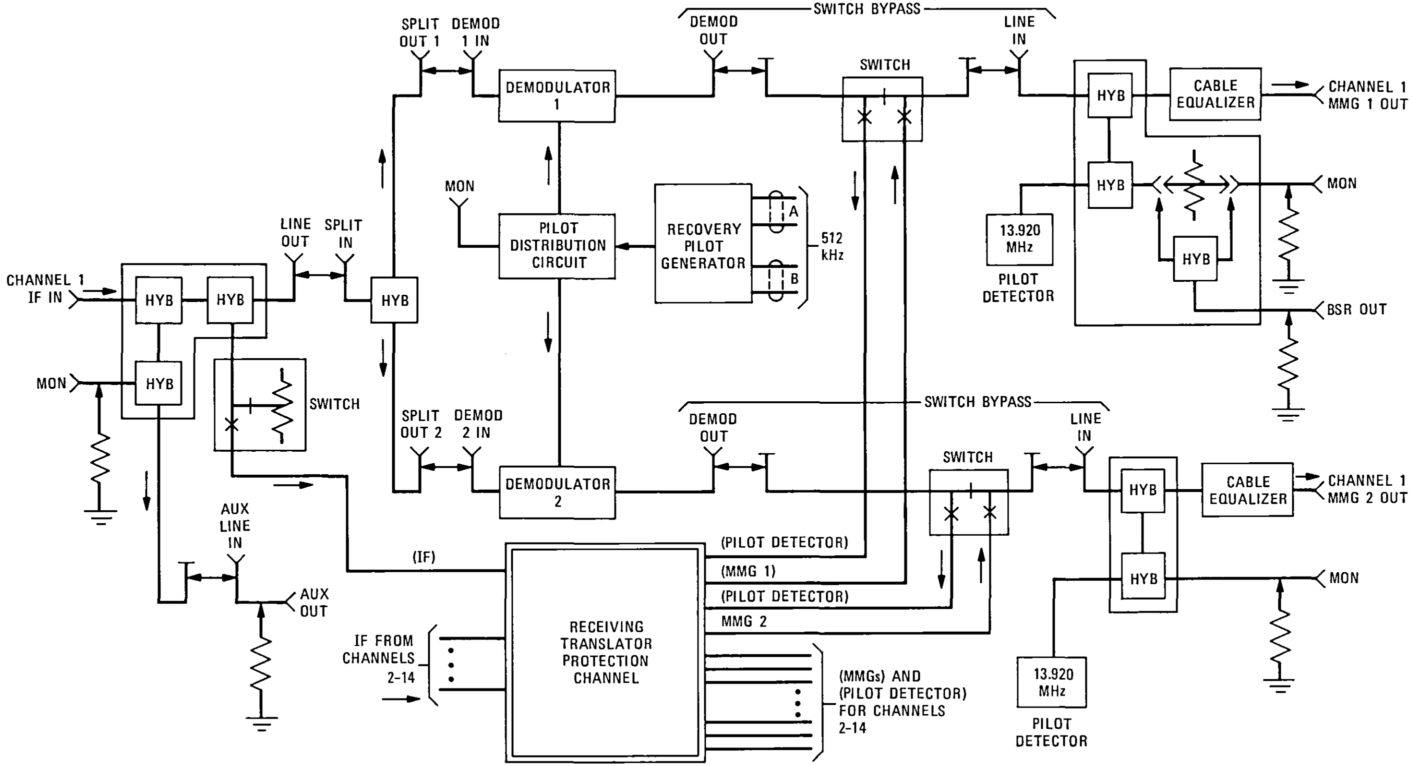

Fig. 5—Simplified block diagram of receiving MMGT-R translator.

frequency control at the receiving station. This function is discussed in greater detail later.

On occasion, it is necessary to use AR6A to transmit the Bell System reference frequency generated at Hillsboro, Missouri, to points not served by other carrier facilities. A BSRTU generator converts a 2.048-MHz reference frequency to two pilot tones separated by 64 kHz. These 11.200- and 11.264-MHz tones are added to MMG1 prior to translation in the MMGT-R transmitter. They are recovered at the receiving MMGT-R where a BSRT receiver converts their difference frequency to the original 2.048-MHz reference frequency. Although the absolute frequencies of the two tones may have shifted because of frequency error attributed to radio repeaters, their precise 64-kHz difference is preserved. Thus, the Bell System reference frequency integrity is preserved.

In the receiving direction, the IF signal from AR6A radio is passed through a splitting hybrid. A pair of upper and lower IF sideband filters separates the signal into its two original sideband components. Each sideband is then translated downward to the multimastergroup spectrum for further demodulation to the basic mastergroup spectrum in MGTB equipment.

3.2.2 Protection terminals

As shown in Figs. 4 and 5, protection transmitting and receiving translators are provided for equipment protection, maintenance, and restoration. The equipment required for this includes switches, pilot detectors, pilot oscillators, and a microprocessor control circuit. The switches, pilot detectors, and oscillators are located in the MMGT-R shelves. The 500B protection switch microprocessor control unit is located in an adjacent bay. Thus, only dc control leads are required between the 500B and MMGT-R bays, and transmission cable lengths are minimized.

Service is routed through the protection translator when a fault is detected in a regular translator or when a manual switch is requested. One protection translator provides protection for up to 14 regular translators. Receiving and transmitting translators are protected independently of each other. This results in two separate 1 × 14 switching arrangements under the common control of a single 500B controller at each MMGT-R location.

Loss of the 13.920-MHz continuity pilot in an MMG is detected by the MMGT-R equipment. When a failure occurs, the 500B controller transfers both MMG1 and MMG2 to protection even though a failure may have been associated with only one MMG. Local and remote alarms and indications are provided by the 500B.

3.3 Detailed description

3.3.1 Transmitting translator

Each transmitting translator includes all of the networks and circuits required to translate a pair of MMG signals to the lower and upper IF frequencies (see Fig. 1). The protection transmitting translator differs from the regular one only in the peripheral circuits associated with the input and output functions. Plug-ins MODULATOR 1 and MODULATOR 2 (see Fig. 6) each contain a balanced diode ring modulator, amplifiers, and a free-running, crystal-controlled carrier oscillator. IF bandpass filters select the lower sideband of each modulator.

A 16.608-MHz recovery pilot generator (RCVRY PLT GEN) is phase locked to a local 512-kHz signal derived from the OMFS which is locked to the Bell System reference. A hybrid transformer network distributes the recovery pilots to the MMGT-R modulators.

The radio-line pilots are generated by three free-running crystal oscillators at frequencies of 11.232, 19.296, and 21.966 MHz and are added to the MMG signals by the hybrid transformer network above.

The distribution network maintains at least 85 dB of isolation between the two MMG signal paths of an MMGT-R.

MMG1 and MMG2 input networks provide terminated monitoring access and split the signal to provide inputs to the protection translator. The Bell System Reference (BSR) input option is required only for MMG1.

An output port is available to make terminated measurements of the IF output signal. An auxiliary input port may be used for restoration. Pilot detectors monitor the translated 13.920-MHz continuity pilots to initiate a switch of both MMGs to protection when the level of either or both drops 6 dB or more.

3.3.2 Receiving translator

The receiving translator accepts an IF signal from AR6A via the 500A protection switch. This IF signal is split into two paths by a hybrid, following which the upper and lower IF bands are selected with IF bandpass filters. Each is then demodulated to the MMG

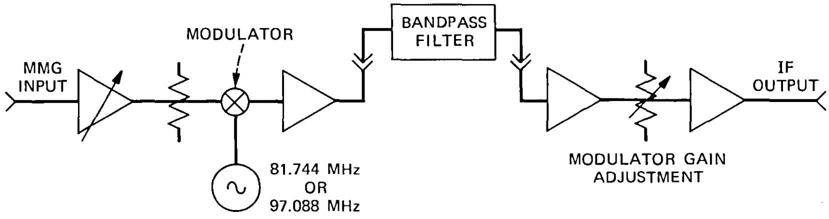

Fig. 6—MMG1 or MMG2 modulator.

spectrum for connection to receiving MGTB equipment. Plug-ins DEMODULATOR 1 and DEMODULATOR 2 (see Fig. 7) use ring-type modulators as in the transmitting direction. The demodulator carriers are generated by voltage-controlled crystal oscillators. Their frequency is controlled by a phase-locked technique to correct the frequency error that accumulates along the radio line. This technique is described below.

A 16.608-MHz recovery pilot is generated in the receiving office translator in a manner identical to that used in the transmitting office translator. It is phase locked to the Bell System reference via the local 512-kHz reference frequency at the receiving site. This 16.608-MHz signal serves as the reference input to the phase-locked loop of each demodulator carrier generator.

From each demodulated MMG in the receiving MMGT-R, the recovered 16.608-MHz pilot, with any frequency offset acquired over the radio line, is selected by a narrowband crystal filter. The phase of this received pilot is compared to that of the locally generated 16.608-MHz signal. Any phase difference between them is converted to a dc control voltage change. This voltage change alters the frequency of the carrier oscillator in a direction to correct frequency offsets of at least ±2500 Hz in the received signal. This technique ensures that the recovered MMG signal is a faithful reproduction of the original transmitted signal.

To enhance the pull-in range of the phase-locked loop, a low-frequency (≈2 Hz) oscillator has been added to the loop filter feedback circuit. This oscillator is activated upon interruption of one or both phase comparator input signals, and when activated, it modulates the

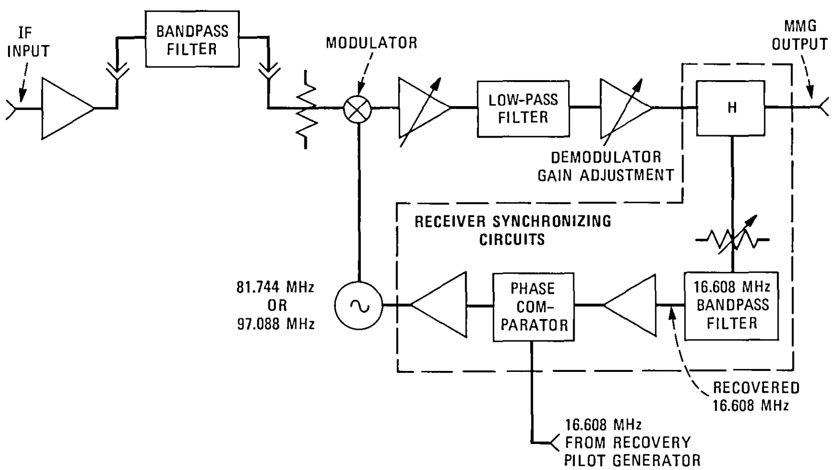

Fig. 7—MMG1 or MMG2 demodulator.

oscillator dc control voltage with sufficient amplitude to sweep the carrier frequency over a range of ±2500 Hz from nominal. This facilitates recapture of a restored signal. In addition, the low-frequency signal is peak detected to generate an out-of-lock alarm and a protection-switch request. Once the loop reacquires lock, the low-frequency oscillator is immediately disabled and has no effect on normal operation.

To accommodate space-diversity applications, the receiver phase-locked loop has been designed such that large phase hits coupled with sudden signal up fades will result in less than 2 ms of loop-phase instability.

The receiving IF network provides terminated monitoring access and splits the signal for protection switching. It also provides an auxiliary output for restoration. Signal monitoring access is provided at both translated MMG outputs. Pilot detectors monitor the recovered 13.290-MHz continuity pilots to initiate protection switching. On a 6 dB or more drop in level of either or both MMG pilots, both MMGT-R translators will switch to protection.

3.3.3 Recovery pilot generator

The 16.608-MHz recovery pilot generator is shown functionally in Fig. 8. Reproducing the transmitted signal accurately is highly dependent upon synchronizing the pilot generator to accurate 512-kHz reference frequency signals. These are derived from the OMFS, which, in turn, is locked to the Bell System frequency source at Hillsboro, Missouri.

To enhance overall reliability, redundant 512-kHz inputs, designated A and B, are supplied to each recovery pilot generator. This is done over separate 135-ohm balanced cable buses and, where possible, via different cable ducts.

The recovery pilot generator accepts two 512-kHz signals, filters them in narrowband crystal filters, amplifies them, and via a divider, converts each to a 32-kHz square wave. A logic switch, preferential to the A signal, provides the selected 32-kHz square wave to one input of a digital phase comparator. The second input to the phase comparator is derived from a portion of the 16.608-MHz crystal oscillator output, which is converted to a 32-kHz square wave. Phase comparison of these two signals results in a dc control voltage that locks the oscillator precisely at 16,608,000 Hz.

Manual test operation of the 32-kHz logic switch is provided. However, phase hits could occur due to reference frequency signal switching. To prevent this, manual switching in the regular channel recovery pilot generator is inhibited unless service is being carried on the protection channel. Conversely, a manual switch in the protection

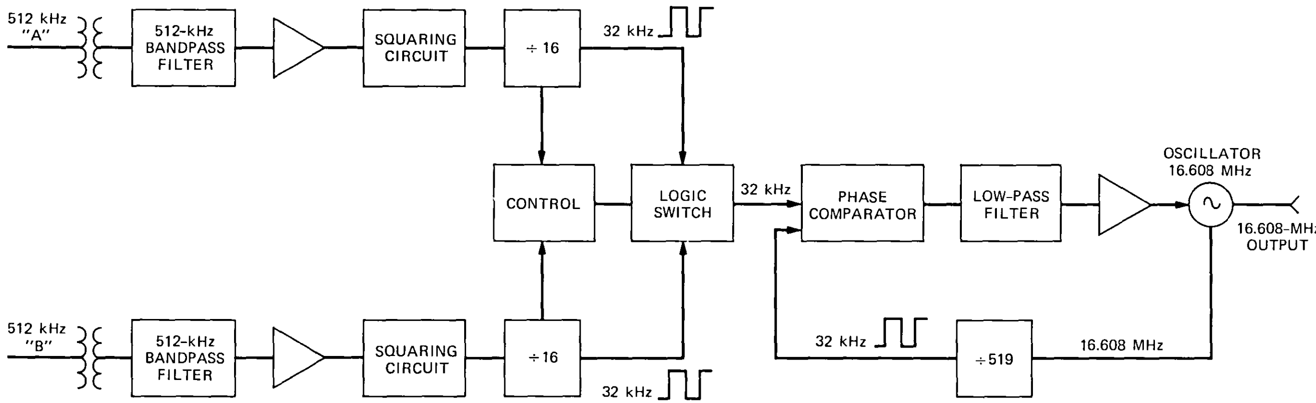

Fig. 8—Functional diagram of a recovery pilot generator.

channel recovery pilot generator can only be made when the protection channel is idle. (See Tables II and III for characteristics and performance of MMGT-R.)

3.4 Realization of objectives

The performance objectives listed represent acceptable levels of performance. Measured data reflect a level of performance better than the values shown. Examples may be seen in Figs. 9 and 10. Figure 9 shows a plot of amplitude versus frequency for a production transmitting and receiving translator pair connected back to back. Figure 10 shows representative noise load data for the same equipment in the same configuration. As the figure indicates, noise performance is thermally limited.

The overall amplitude versus frequency curve in Fig. 9 is a nonadjustable parameter. It is influenced by the response of the individual components (amplifiers, filters, etc.). The only equalization used with

Table II—Characteristics and performance of MMGT-R

Characteristics	Performance
Type of modulation	SSB-SC-AM
Number of MMGs	2 per AR6A channel
Number of MGs	5 per MMG (10 total)
MMG spectrum (MG 3-7)	8.628 through 21.900 MHz
IF output spectrum	59.844 through 88.460 MHz
MMG transmit input level	−41.6 dBTL
IF transmit output level	−23.9 dBTL
IF receive input level	−32.1 dBTL
MMG receive output level	−22.4 dBTL
Return loss (all ports)	>25 dB
dc power—input voltage	−20 to −28 volts dc
dc current drain (per shelf)	2.8 amperes maximum at −24 volts dc
Normal power dissipation per 7-foot bay (3T/R pairs)	400 watts
Protection	500 B − 1 × 14 maximum

Table III—Performance of MMGT-R

Performance Objectives	Worst-Case Deviation
Gain frequency response (1 − TR pair)	<±0.3 dB peak to peak across any MG
Noise performance (equivalent 6000-circuit load, 1 − TR pair)	<15 dBrnc0
Crosstalk	<−100 dB
Carrier leak	⩽−70 dBm0
Spurious tones	⩽−70 dBm0
Phase-jitter performance for ten terminal pairs in tandem	⩽1° peak to peak, Bell weighted
Frequency stabilities (over 80° ±20°F temperature range and permissible power supply variation)	
Transmitter carrier oscillator frequency	±200 Hz
Radio-line pilot oscillator frequency	±4 parts per million

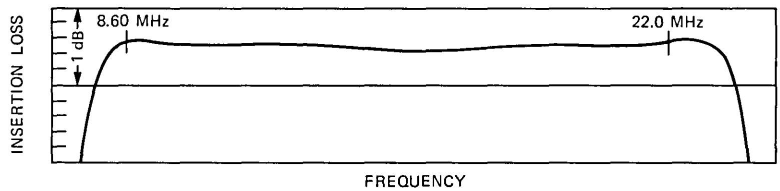

Fig. 9—Typical amplitude vs. frequency response of transmitting/receiving translator.

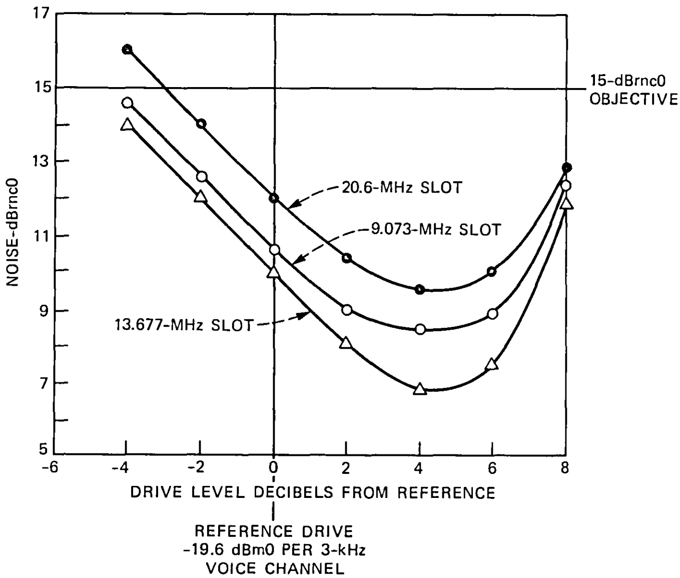

Fig. 10—Typical noise load performance of transmitting/receiving translator.

MMGT-R compensates for amplitude slope due to cable lengths in trunks between the MMGT-R, the MGTB, and the 500A equipment.

Mastergroups assigned to AR6A contain a mixture of analog-voice and voiceband data circuits subject to impairment due to excessive phase jitter. Strict phase-jitter requirements are placed on the transmit and receive carrier generators and on the recovery pilot generator to minimize phase-jitter transfer to data signals.

3.5 Physical design considerations

3.5.1 Goals

The physical arrangement of the MMGT-R equipment is oriented to achieve three major goals: (1) high reliability; (2) ease of installation, maintenance, and growth; and (3) cost-effectiveness. The first two items influence the third.

3.5.2 Reliability

When considering that an MMGT-R shelf carries up to 6000 two-way voice circuits and that a 7-foot bay carries up to 18,000 such circuits, reliability is of utmost concern. With this in mind, an early decision was made to minimize the amount of common equipment and to have each transmitting and receiving translator function independently.

This simplified the protection aspects of the system and minimized the risks to remaining operational parts of the system when equipment failures require repair or maintenance.

Carriers and recovery and radio-line pilots are all generated on a per-shelf basis with plug-in modules that may be replaced easily.

Converters of the dc-to-dc type are provided on a per-shelf basis with separate power for switching and transmission functions. A switching power failure does not impair transmission, and active transmission circuits are fully protected.

By locating switches and detectors in the same shelf as the equipment they protect, transmission cable lengths and interbay wiring and cabling are minimized.

The shelf and plug-in layouts promote rapid and natural heat dissipation. A combination of baffles and perforated cover assemblies permit the free flow of air vertically and from front to rear of each shelf. Each shelf has its own airflow pattern, avoiding high concentration of heat in the upper region of the bay.

3.5.3 Ease of installation, maintenance, and growth

The combination of the factors described previously led to a completely self-contained, totally connectorized shelf design. Thus, it requires no installer wiring for initial or subsequent growth installation. An alarm and fuse panel at the top of each bay constitutes the only MMGT-R common equipment. The protection MMGT-R and switch controller are naturally required with the initial installation to permit subsequent growth.

Redundant diode-firmed −24 volt power leads and 512-kHz reference frequency leads are connectorized and provided from the fuse panel to each shelf position in the bay. All active circuitry is of the plug-in type and all passive circuitry uses connectors for ease of removal or replacement. Access ports are provided for components outside protection switching boundaries to readily permit their bypassing for removal or replacement without delaying service restoration. These design concepts achieve a highly reliable and easily maintained circuit and equipment arrangement capable of "graceful growth."

3.5.4 Cost-effectiveness

As we stated earlier, achievement of the first two goals contributes significantly to providing a cost-effective system. By minimizing the amount of common equipment, the costs for new installations or for low cross-section systems with limited growth potential are reduced. Thus, the costs per circuit mile tend to be distributed more evenly. The reduced time required for installation and maintenance also contributes to lower overall operating expenses as well.

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Antoine Dubois, E.E. Assoc., 1968, Merrimack College; Western Electric 1960–1967; Bell Laboratories, 1967—. Mr. Dubois began working at Western Electric Quality Assurance in 1960, and in 1967 joined Bell Laboratories, where he participated in the lineup of line-connecting and mastergroup-multiplexing circuits along the initial L-4 route. Since then, his responsibilities expanded to mastergroup administration. This includes mastergroup-multiplexing and associated circuitry used for multimastergroup transmission in cable and radio systems.

Donald N. Ritchie, Certificate (Mechanical Engineering), 1941, Stevens Institute of Technology; Western Electric, 1941–1948; Bell Laboratories, 1948—. While at Bell Laboratories Mr. Ritchie has been involved with design standards, short-haul radio system applications, outside supplier auxiliary channel radio applications, data under voice, multiplex terminal physical design, and pulse-code modulated terminal physical design.

Frederick M. Smith, A.E.E., 1960, Wentworth Institute; Bell Laboratories, 1960—. Since starting with Bell Laboratories Mr. Smith has been involved with such projects as short-haul microwave radio and terminal facilities. He is presently responsible for the MMGT-R multiplex terminal used with AR6A single-sideband radio systems.

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The AR6A Single-Sideband Microwave Radio System:

Frequency Control

By J. M. KIKER, Jr.* and S. B. PIRKAU*

(Manuscript received February 14, 1983)

The AR6A Radio System requires precise control of the local oscillator frequency at repeaters and main stations, so that the IF equalizer pilots will be accurately located with respect to the narrowband crystal pick-off filters. In addition, the accumulated frequency error from each repeater and main station must be limited to a value that can be corrected by the tracking receiver of the multimastergroup terminal. Precise frequency control is maintained along the route by locking local oscillators to a stable and accurate frequency reference provided at each repeater station by the Microwave Carrier Synchronization Supply (MCSS). The MCSS provides a source and a backup for this signal. Extensive monitoring circuitry in the MCSS detects failed conditions. The accuracy of the two oscillators is continuously monitored by making frequency comparisons of their output. These frequency-monitoring circuits are also used to make yearly adjustments to the oscillators using a rubidium frequency standard as reference. This paper describes the MCSS System that has been developed for the AR6A Radio System.

I. INTRODUCTION

Compared with current 6-GHz FM Radio Systems like TH-1 or TH-3, the new AR6A† Radio System requires much tighter radio-frequency tolerances to keep radio-line pilots within narrowband pick-

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

off filters. These pilots should remain within ±2 kHz of the center frequency of their pick-off filters, which translates to a frequency tolerance of about three parts in 10^7 compared to the TH-3 tolerance of two parts of 10^5. This tighter tolerance is beyond the capability of currently available microwave generators used as carrier sources for up conversion and down conversion in the radio Transmit-Receive (TR)* units. The required stability is obtained by synchronizing the microwave generators and shift oscillators to a highly stable frequency synchronization supply located in each radio station. This AR6A Microwave Carrier Synchronization Supply (MCSS) is described in the following sections. The paper then presents an overall block diagram description of the MCSS and discusses in more detail the various subsystem components.

II. FUNCTIONAL DESCRIPTION OF THE MCSS

The MCSS distributes a highly accurate sine-wave frequency of 308.8735416 kHz (abbreviated as 308.9 kHz) to each TR-bay microwave generator and shift oscillator as illustrated in Fig. 1. All microwave-generator frequencies and the shift-oscillator frequency are harmonics of this reference signal and are locked to it by phase-locked loop circuitry located in the frequency control units. The distribution bus provides 32 isolated outputs that can supply reference signals to a fully loaded route of eight two-way AR6A radio channels (eight channels by two directions by two oscillators per bay).

Figure 2, a block diagram of the MCSS, shows two symmetrical sections (frequency supplies), one of which is used for on-line operation and the other for standby. The major component in each of the two sections is a crystal-controlled, 4.94197666-MHz (hereafter abbreviated to 4.94 MHz) oscillator that has a long-term frequency stability of 1.8 parts in 10^7 per year. Each oscillator is associated with a divider/switch unit, which performs both a frequency division of 16 and power monitoring functions. Although only one of the two oscillators is used at a time, a continuous frequency comparison between them checks for an excessive frequency difference and generates appropriate alarm signals if the difference exceeds certain values. The interconnection between the two sections is accomplished through the control unit, which offers automatic switching between output signals under trouble conditions or manual switching for maintenance purposes.

Figure 3 is a picture of the MCSS that shows the major component parts described above.

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

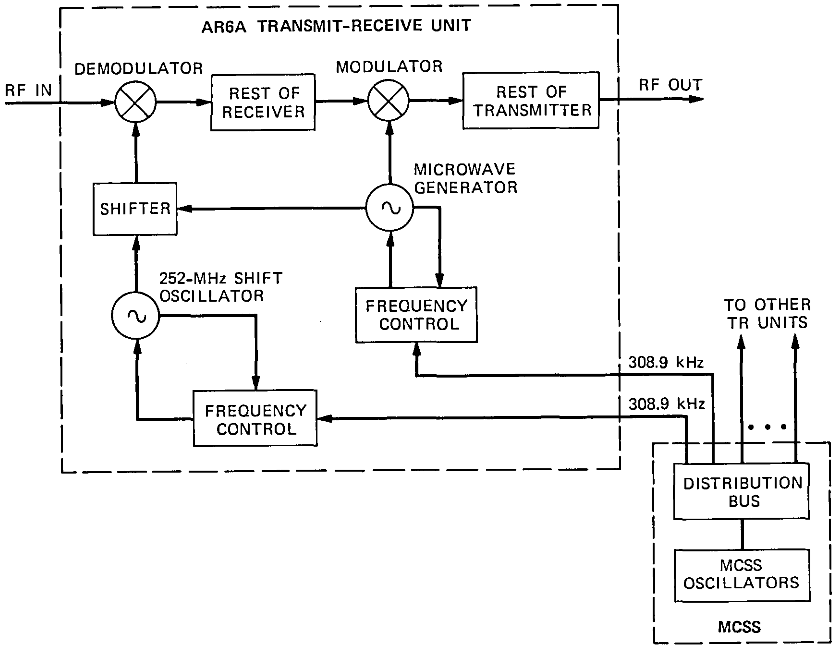

Fig. 1—Synchronization of AR6A TR units to MCSS.

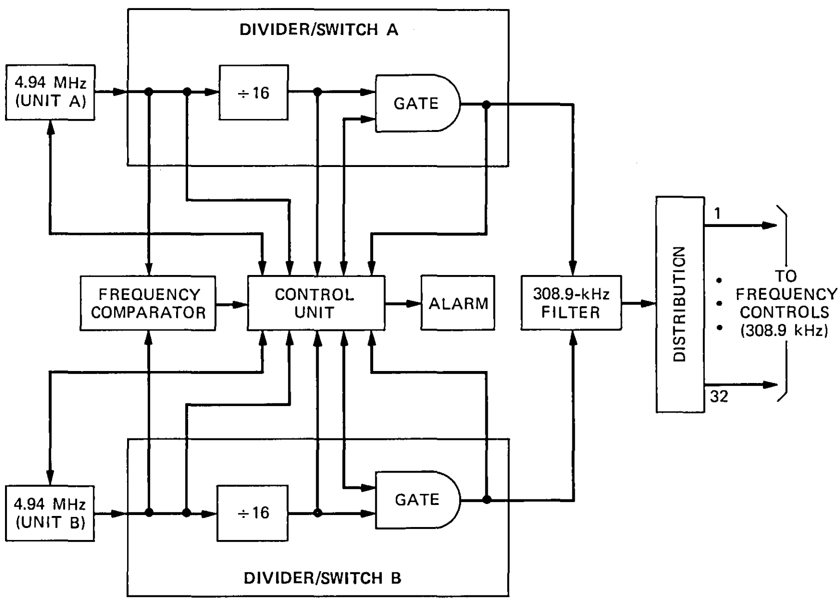

Fig. 2—Block diagram of MCSS.

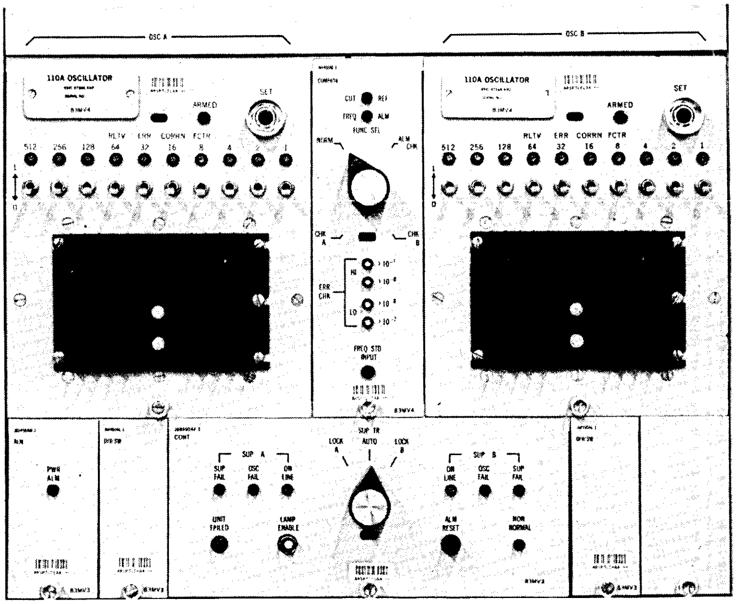

Fig. 3—Actual MCSS.

III. DETAILED CIRCUIT DESCRIPTION

*3.1 4.94-MHz oscillator**

The 4.94-MHz oscillator, shown in block diagram form in Fig. 4, consists of a crystal-controlled oscillator, RF amplifier, frequency-controlling network, and temperature-control network in a thermally stable environment. Because of repeater-station temperature variations, use of a stable oven is necessary to obtain the required frequency stability. Other peripheral circuitry such as the heater power amplifier, the binary frequency-controlling switches, and their display Light Emitting Diodes (LEDs) are exposed to ambient temperature.

3.1.1 Radio Frequency (RF) oscillator and amplifier

The RF oscillator is a modified Pierce-type crystal oscillator using a precision quartz crystal unit. A capacitance and varactor-diode network is in series with the crystal for frequency adjustment of ±3 ppm from the nominal frequency of 4.94197666 MHz. This stage has automatic gain control to ensure stable frequency operation. The crystal unit is a third-overtone AT-cut resonator having a turnover temperature between 76 degrees and 86 degrees Celsius. The output

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* The MCSS oscillator was designed by M. W. Zuidervliet, Jr. of Bell Laboratories in Allentown, PA.

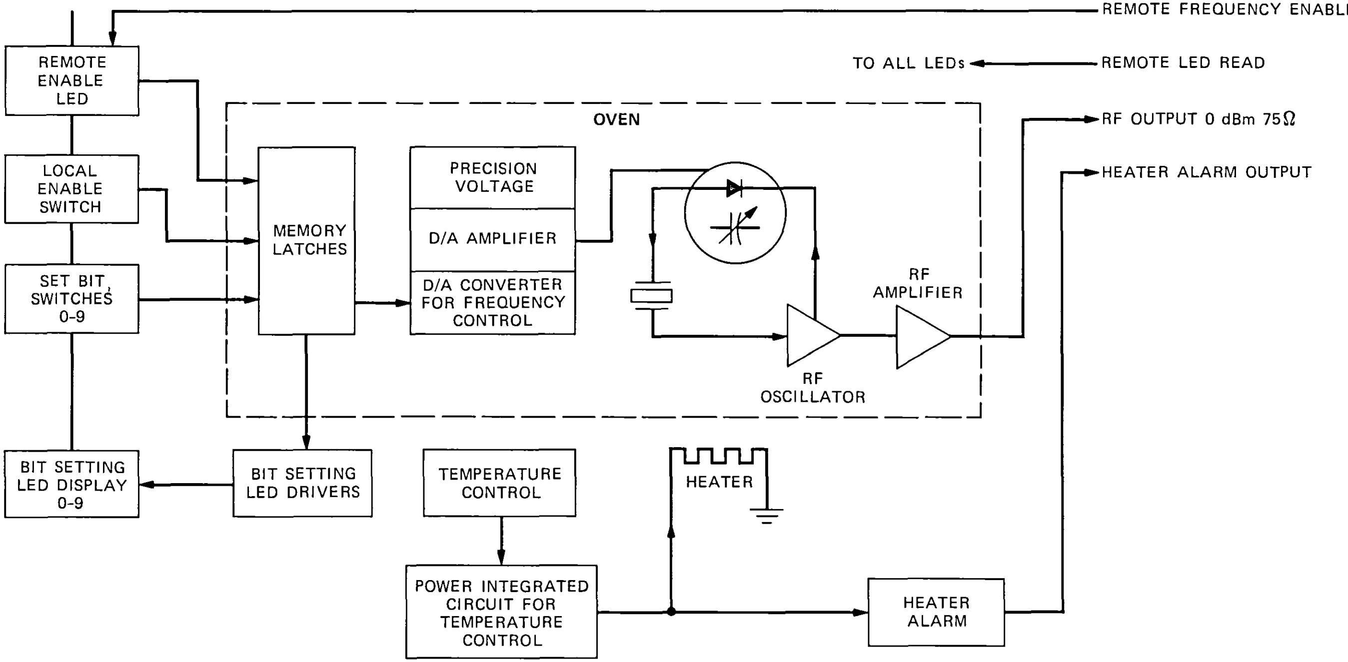

Fig. 4—Block diagram of MCSS oscillator.

of the oscillator is at the base of a low-noise transistor that uses the crystal unit to filter out any noise generated in the transistor that is more than a few hertz from the operating frequency. The output impedance of the oscillator is 50 ohms.

The RF amplifier consists of four transistor stages that buffer the oscillator from the output and boosts the output signal level of 0 dBm at an output impedance of 75 ohms.

3.1.2 Frequency tuning circuitry

The varactor diode, in series with the crystal, is biased from the frequency adjust circuitry made up of a precision voltage reference source, Digital-to-Analog (D/A) converter, and current-to-voltage operational amplifier. Front-panel toggle switches control memory-latch circuits that, when enabled, supply digital input to the D/A converter. The D/A converts this input to a precision current value utilizing the precision voltage reference. A voltage proportional to this current is applied as bias to the varactor diode. The ten toggle switches are arranged for binary weighting with the least-significant bit giving a frequency change of one part in 10^8. A new switch setting is input to the memory latches only when a remote enable signal is present and a load button on the oscillator is depressed. At other times the position of the toggle switches has no effect. This feature is present to prevent accidental detuning of the oscillator. Since the switch settings do not necessarily agree with the status of the memory latches, display LEDs are provided on the front panel that, when enabled, indicate the value of the digital input to the D/A.

3.1.3 Temperature control and heater alarm

The temperature of the oven is maintained by a proportional temperature control using a Wien Bridge oscillator where the sensing elements are two thermistors in series. These thermistors are imbedded in the wall of the oven and are tightly coupled to the heater winding. The output of the Wien Bridge oscillator is buffered by an amplifier that has variable gain for adjustment of the sensitivity of the temperature-control circuit. This amplifier drives a power integrated circuit that provides the dc power to the heater winding. The power stage limits the current into the 15-ohm heater winding to approximately 1-1/3 amps.

There is an alarm circuit that senses the heater-winding control voltage and emits an alarm if this voltage moves beyond the window for normal operation. This is an indication that the oscillator already has or is about to lose temperature control and therefore change frequency.

3.2 Divide-by-16 circuit

The main purpose of this circuit is to convert the 4.94-MHz input to 308.9 kHz and gate this signal to the distribution bus when the supply is on-line. There are two identical circuits corresponding to sides A and B.

Figure 5 shows a block diagram of the divide-by-16 circuit. Following the main signal path, the 4.94-MHz sinusoidal input signal is converted to a Transistor-Transistor Logic (TTL)-compatible square wave and divided by 16. The signal from the divider is input to one side of an AND gate. If this supply is to be selected as on-line, a select signal from the control unit enables the other gate input and the 308.9-kHz signal is output to the distribution bus.

Various monitors in the circuit detect improper operation. At the input, a detector monitors the level of the incoming 4.94-MHz signal and, if that level drops 2 dB, outputs an alarm to the control unit. A gradual 2-dB drop in level is not sufficient to cause the sine-wave-to-square-wave comparator to fail; thus, the alarm to switch to the other supply can be given before actual loss of 308.9 kHz occurs.

A coarse frequency monitor of the 308.9-kHz signal guards against a failure of the divider. This will detect a gross error that would occur if the divider malfunctioned to produce the wrong division ratio. A detected error of this type is output as an alarm to the control unit.

The level of the 308.9-kHz signal output to the distribution bus is monitored. The output gate for this signal is the last active circuit component of the reference signal path to the TR-bay frequency control units. The alarm indication for loss of level is output to the control unit but is meaningful only if the gate output is on-line.

The circuit buffers and provides a TTL line driver output of the 4.94-MHz signal to the frequency comparator unit where precise frequency comparisons between the two oscillators occur.

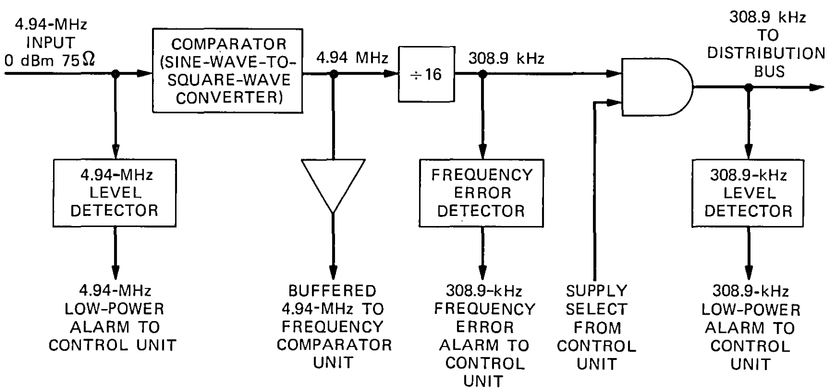

Fig. 5—Block diagram of MCSS divide-by-16 circuit.

3.3 308.9-kHz distribution

The outputs of the two divider switch circuits are gated to the distribution circuit. Either circuit, but only one at a time, provides a 308.9-kHz TTL signal for the distribution bus. The two signal inputs are combined in a resistive power combiner followed by a 308.9-kHz bandpass filter that provides at least 40 dB of loss to frequencies of 618 kHz and higher with a passband loss of less than 1 dB.

The 308.9-kHz sine-wave output from the filter, at an impedance level of 135 ohms, is fed to a balanced transformer with a 135- to 8.06-ohm impedance transformation. This output drives a passive resistive distribution bus that provides 33 outputs balanced at an impedance of 135 ohms. Thirty-two of the outputs are for distribution to the TR bays and are terminated if not used. The remaining tap provides a monitor port for measurement of signal level at the bus output. There is 18 dB of signal loss from the filter output to any tap output, and the tap-to-tap isolation is 36 dB.

3.4 MCSS control unit

The interconnection between the two 308.9-kHz sources of the MCSS is accomplished through the control unit. The control unit handles all MCSS alarms, as well as power and frequency sensor functions, and performs four separate major functions:

1. Selects (puts on-line) one of the two MCSS oscillators as the 308.9-kHz source.
2. Offers automatic switching to the other MCSS oscillator should the on-line unit develop any trouble.
3. Originates the general MCSS alarm together with visual indications pinpointing the specific alarm condition.
4. Furnishes the interface for the two oscillators, two divider and switch units, the frequency comparator unit, and the alarm unit.

All of the above-listed functions can be performed in either the auto or manual operating modes. The auto mode of operation is the normal operating condition of the control unit. In this mode, all sensors are monitored and an automatic switch will occur if there is a failure in the on-line supply. The sensors monitor the following items:

1. Power output of the 4.94-MHz oscillators
2. Operation of the oven within the oscillator unit
3. 308.9-kHz coarse output frerquency and power
4. Frequency difference between the two oscillators.

The sensors for items 1 and 3 are located on the divider and switch units and for item 4 on the frequency comparator unit.

The most severe alarm condition is the cut-reference alarm. When this alarm is generated the MCSS outputs to the TR bays are removed and the microwave generators and shift oscillators become free running. This alarm can be generated for two reasons:

1. The frequency offset between the two oscillators is greater than 39.3 parts in 10^8 (measured by the frequency comparator unit).

2. A failure has occurred first in the on-line section, causing an auto switch, and then in the former standby section, which was put on-line by the auto switch mentioned above. If the failure is in the standby supply, an oscillator fail alarm is generated and all switching between supplies is inhibited. Subsequent failure of the on-line supply will cause a cut-reference alarm.

The manual mode of operation is primarily for maintenance and repair of the MCSS. It permits a manual switch resulting in an interchange of the on-line and the standby section of the MCSS. However, consideration had to be given to the fact that whenever a manual switch to the standby supply is executed, a check has to be performed to determine if the newly selected supply does have a signal output of correct amplitude. Therefore, once a new supply is selected and on-line, enough time has to be allowed to permit a switchback to the previous on-line supply should the new one fail to produce a suitable output signal. Delay circuits permit this.

Since an auto switch will only occur as a result of a failure in the on-line section of the MCSS, a switchback is not possible and the check for a good output signal of the standby section is not necessary.

3.5 MCSS comparator unit

3.5.1 Functional description of the MCSS comparator unit

The comparator unit performs two major functions in the MCSS.

1. For normal operation, there is continuous monitoring of the frequency difference between the two 4.94-MHz oscillators. An alarm signal is generated if the frequency difference exceeds 16.6 parts in 10^8. A second alarm and cut-reference signal is generated when this difference increases to, or exceeds, 39.3 parts in 10^8. The cut-reference signal removes the synchronization tone from the microwave generators and shift oscillators, and these units become free running.

2. Replacement or yearly routine maintenance requires frequency adjustment of the 4.94-MHz oscillator. In the frequency check mode of operation, the comparator aids in oscillator adjustment by indicating the frequency difference between the oscillator being adjusted and an external frequency calibration standard. In this mode, the comparator has a resolution of ± 1 part in 10^8, which is based on the settability of the 4.94-MHz oscillators whose smallest frequency adjustment increment is one part in 10^8.

The position of the front-panel mounted function select switch, which can be in either the check A or check B position, determines which 4.94-MHz oscillator is being checked against the external frequency standard. In the check B position, the external frequency standard takes the place of oscillator A and in the check A position the external frequency standard takes the place of oscillator B.

When in the check A or check B position, all frequency alarms are inhibited. The four light-emitting diodes on the front panel of the unit are enabled. These LEDs are labeled and arranged in the following manner:

10^{-7}
HIGH
10^{-8}
10^{-8}
LO
10^{-7}

The HIGH or LO indicates the direction of frequency deviation of the checked 4.94-MHz oscillator in comparison with the frequency standard. By observing the LEDs, the oscillator can be adjusted to within one part in 10^8 of the 4.94-MHz signal.

Adjusting the oscillator is accomplished through ten toggle switches, labeled zero through 9, arranged in a binary weight fashion with the least significant bit being switch zero.

3.5.2 Method of detecting frequency offset

The conventional method of detecting a frequency difference between two oscillators, in which one oscillator will generate a gate pulse during which the other oscillator is counted, results in large time intervals to detect small offsets. For example, to detect one part in 10^8 for the MCSS 4.94-MHz oscillators would require a minimum time of $1/\Delta f = 20$ seconds, which is an unreasonable time interval.

The method chosen for measuring frequency differences in the MCSS is the same as that used on the Jumbogroup Frequency Supply (JFS) of the L5 System, which Fig. 6 shows utilizing the block diagram of the MCSS comparator. In this method, one of the two frequencies being compared is divided by D and then mixed with itself, with the sum signal $A + A/D$ chosen by a bandpass filter. This signal is mixed with oscillator B and passed through a low-pass filter with the resulting signal difference of frequency

$$A + \frac{A}{D} - B = \frac{A}{D} + \Delta f.$$

This signal is used as a clock signal for an N divider that generates a gate pulse. During the gate-pulse interval, counts are accumulated by the C counter. This has the effect of multiplying the real fractional

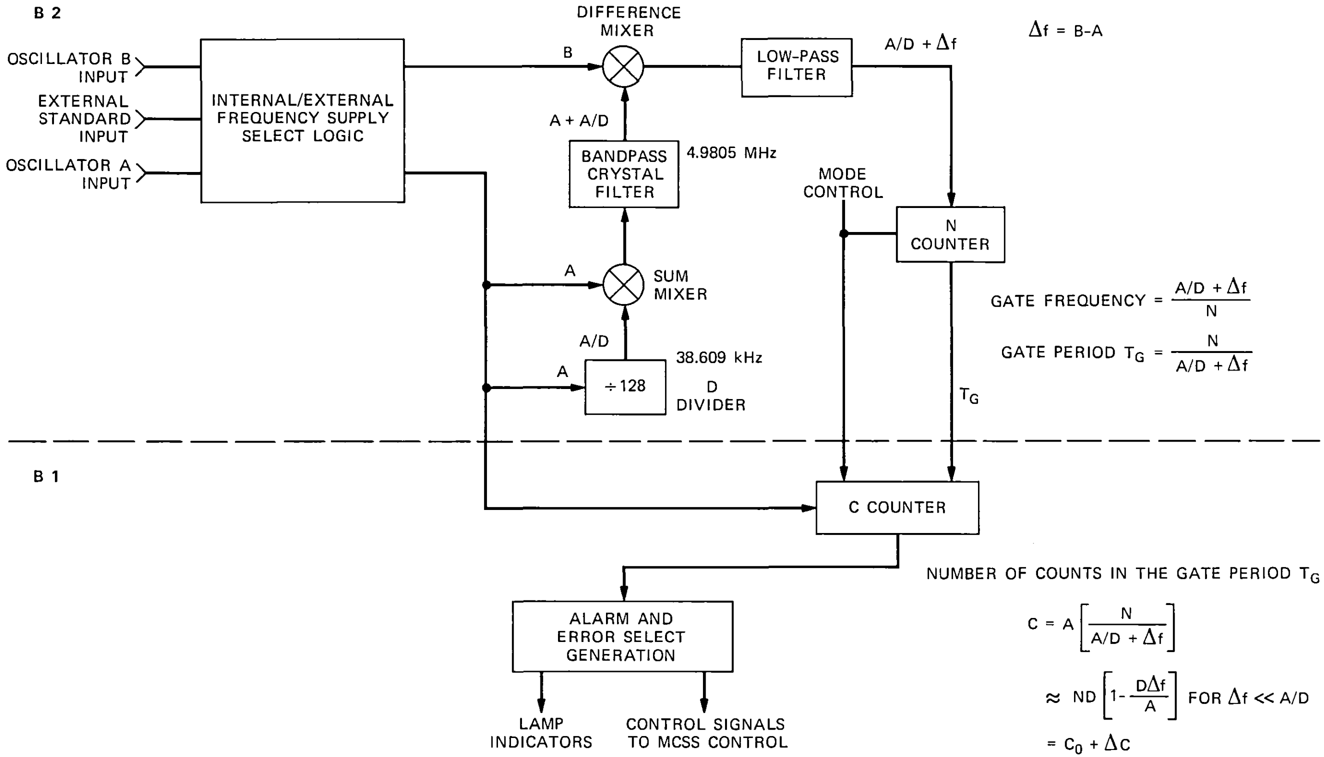

Fig. 6—Block diagram of MCSS comparator unit.

frequency error by a factor D. For the MCSS a value of $D = 128$ was chosen as realistic when considering the need for the bandpass filter that must pass $A + A/D$ and reject $A - A/D$. Now the minimum time to detect an error of one part in 10^8 is the time for $ND^2\Delta f/A$ to equal one, i.e., one extra or one less count than normal for no frequency error. Solving for N, one obtains

$$N = \frac{A}{D^2\Delta f}$$

The gate interval for the minimum count to occur is $T_G \approx ND/A = 1/D\Delta f$. This time interval is shorter than that for the conventional methods by a factor of 128.

3.5.3 Block diagram and description

Figure 6 shows the block diagram representation of the comparator unit. Two printed circuit boards (B1 and B2) are used to accommodate the large number of integrated circuits and the relatively large inductors comprising the low-pass filter. The inputs to the comparator unit, during normal operation, are two 4.94-MHz TTL signals derived from oscillators A and B. For routine maintenance and oscillator calibration, a third input signal, also TTL compatible, from an external frequency standard is also connected to the comparator unit.

The input signal frequency to the 14-stage N counter is $A/D + \Delta f$. The output of this counter is the gate pulse (T_G) for the C counter. The 22-stage C counter is preset to a value C_0 dependent on the operating mode. The clock input signal is from oscillator A, which counts down the C counter during the gate period T_G. If oscillators A and B are identical in frequency, the C counter will count down from C_0 to zero. The value of the C counter is determined by monitoring the Q signals of each counter stage. The Q signals are gated together to detect values of ΔC which correspond to frequency differences. The count can be either positive or negative, depending on whether the counter went through zero or not. For normal operation, only the magnitude of the ΔC count is significant. However, during frequency adjustment of the two oscillators, the sign of the count is also important because the frequency of the oscillator being adjusted is compared to an external frequency standard and the sign of the frequency error must be known for proper adjustment. In the frequency check modes, instead of generating alarms for excessive frequency differences, the comparison circuit uses the four LEDs on the front panel to indicate the magnitude and direction of frequency deviation of the oscillator.

3.5.4 Other design considerations

Except during oscillator adjustment and routine maintenance, the comparator unit operates in the normal mode. In this mode, only a frequency alarm and cut-reference signal and alarm are generated when the frequency difference between MCSS oscillators is 16.6 parts in 10^8 or 39.3 parts in 10^8, respectively. The frequency comparison period in this mode of operation is only 78 milliseconds. It is possible for erroneous counts to occur because of plant battery hits or other office transients. To guard against false alarms occurring due to these erroneous counts, four-stage shift registers are used to output the frequency alarm and the cut-reference signal. To obtain an output from these shift registers, four consecutive readings exceeding the appropriate ΔC limit must occur. This, of course, delays the alarm signals by 312 milliseconds, but is quite acceptable in this application and does eliminate false alarms or, more seriously, loss of sync to an entire station because of an erroneous cut-reference signal.

3.6 MCSS dc power

The MCSS dc power is provided by regulated dc-to-dc converters (−24 to +5 volts dc), which are current limited, over-voltage protected, and low-voltage alarmed. The two 4.94-MHz oscillators are powered by separate 5-volt dc converters. These same converters also individually power the divider/switch circuits. The comparator and control unit are dual fed from the two converters using diodes. These procedures are used in the MCSS to guard against reference signal loss due to failure of any single power unit.

3.7 Alarm unit

The alarm unit in the MCSS provides visual and audible alarms of low power and other trouble conditions in the MCSS. It generates remote indications via telemetry systems to the alarm center, as well as local station alarms. Standard ACO features are available that can be activated at any miscellaneous key, jack, and lamp panel in the support bay or at any TR-bay display panel.

IV. SUMMARY

The paper has described the important features of the AR6A MCSS. The MCSS provides a precise reference signal at 308.9 kHz to which the microwave generator and shift oscillator in a repeater bay are locked. The MCSS is settable to one part in 10^8 and is stable to within five parts in 10^{10} per day. Extensive monitoring of performance and the provision of a standby reference signal ensure a high degree of reliability to the system.

V. ACKNOWLEDGMENTS

The authors would like to acknowledge the efforts of M. W. Zuidervliet, who designed the 4.94-MHz oscillator, and N. B. Rowe and

J. R. Scoville, who contributed greatly in the MCSS design and evaluation.

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AUTHORS

J. Morgan Kiker, Jr., B.S.E.E., 1962, University of California at Berkeley; M.E.E., 1964, New York University; Bell Laboratories, 1962—. Mr. Kiker began work on high-resolution radar systems. Since 1971 he has supervised work on ultralinear IF amplifiers and low-noise carrier and oscillator supplies for the AR6A Radio System. He is now working on digital radio systems. Member, Tau Beta Pi, Eta Kappa Nu.

Siegfried B. Pirkau, A.A.S., 1969, New Hampshire Vocational Technical College (NHVTC); Bell Laboratories, 1969—. Mr. Pirkau's first assignment was work involved with computer-simulated *Picturephone** coding. In 1971 he began work on single-sideband radio in the Repeater Evaluation Group, where he became involved with the design and construction of special test equipment. In 1980 he joined the Digital Radio Laboratory doing DR6 and DR11 development work.

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The AR6A Single-Sideband Microwave Radio System:

Microwave Carrier Supply

By J. GONDA* and J. M. KIKER, Jr.*

(Manuscript received February 14, 1983)

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-

* Bell Laboratories.

† 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

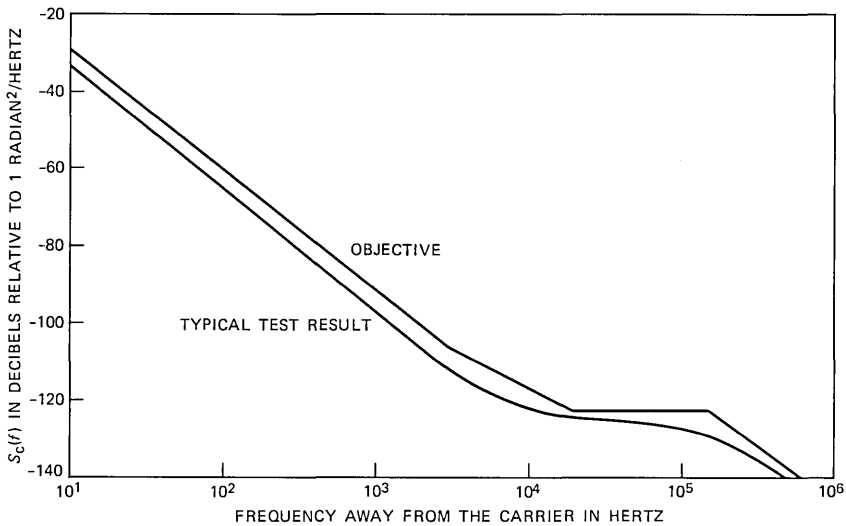

Fig. 1—Phase-noise objective and typical test result for the microwave generator.

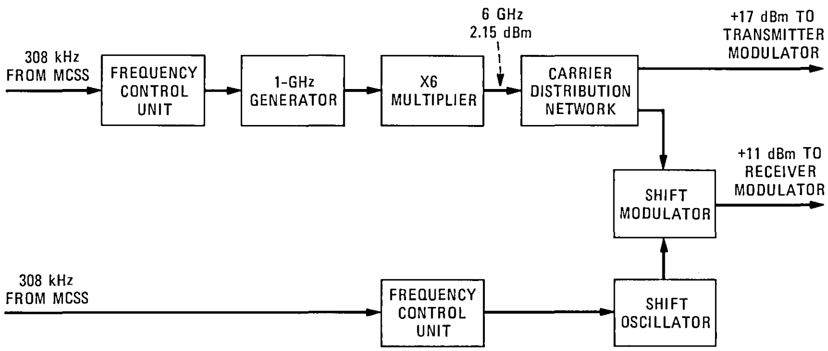

Fig. 2—Microwave carrier supply.

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.

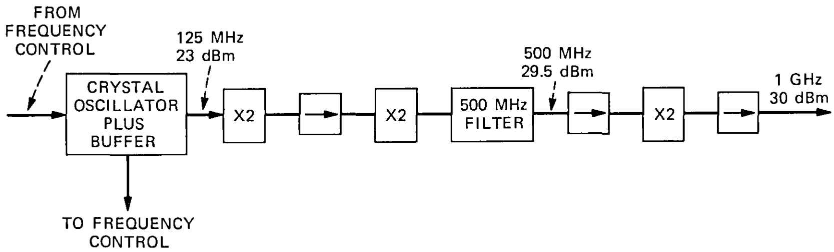

Fig. 3—1-GHz generator.

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.

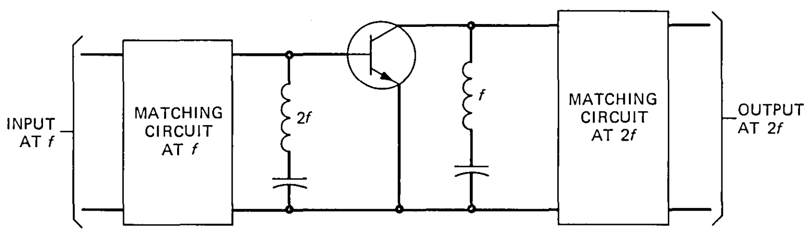

Fig. 4—Transistor doubler.

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

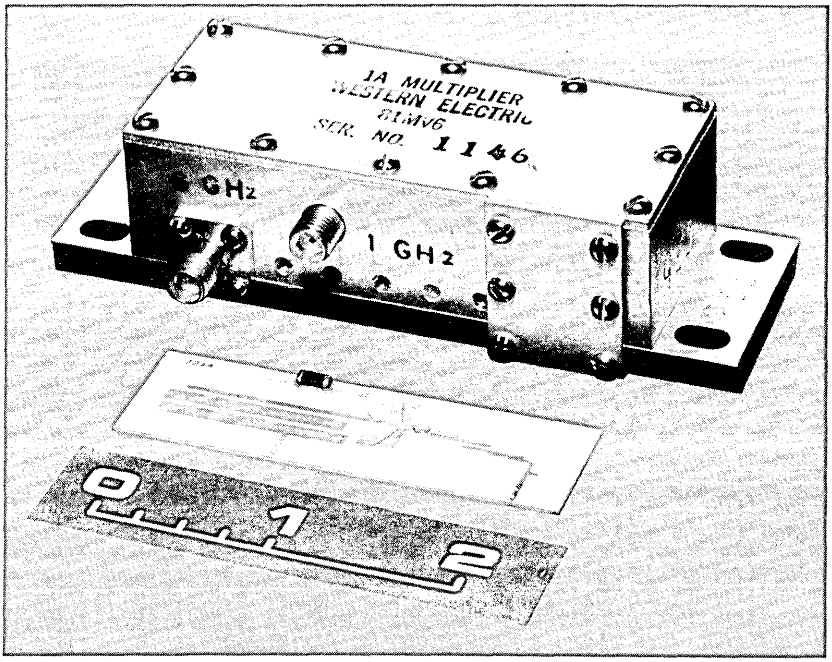

Fig. 5—Photograph of the 6X multiplier.

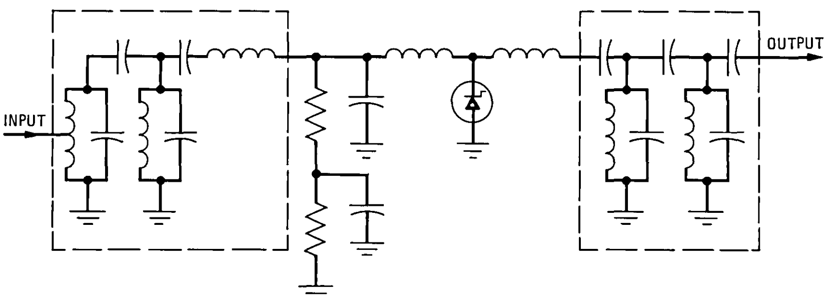

Fig. 6—Lumped element equivalent circuit of the 6X multiplier.

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-

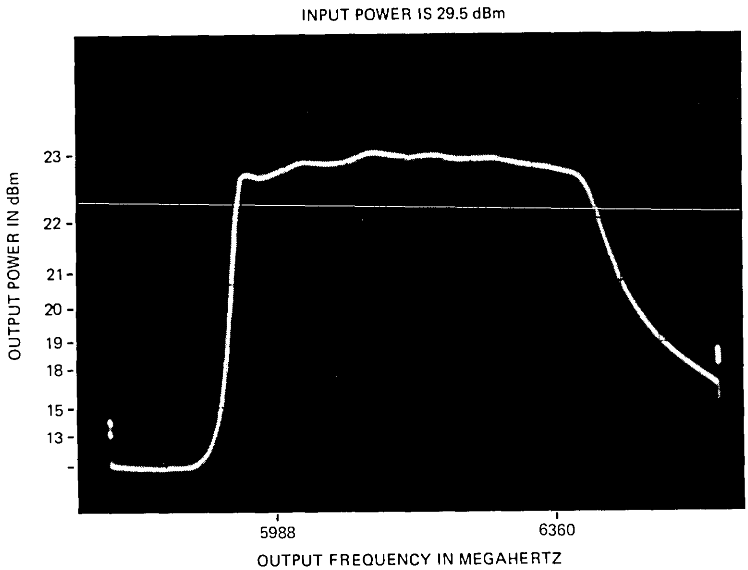

Fig. 7—Typical swept response of the 6X multiplier.

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

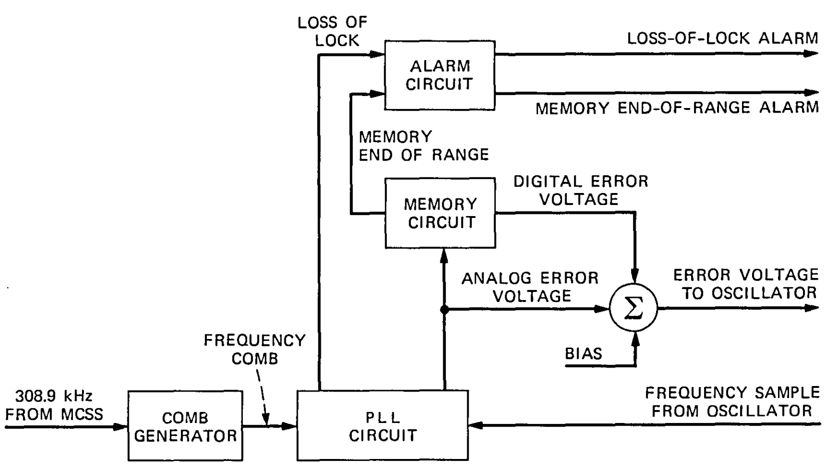

Fig. 8—Block diagram of frequency control unit.

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

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

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.

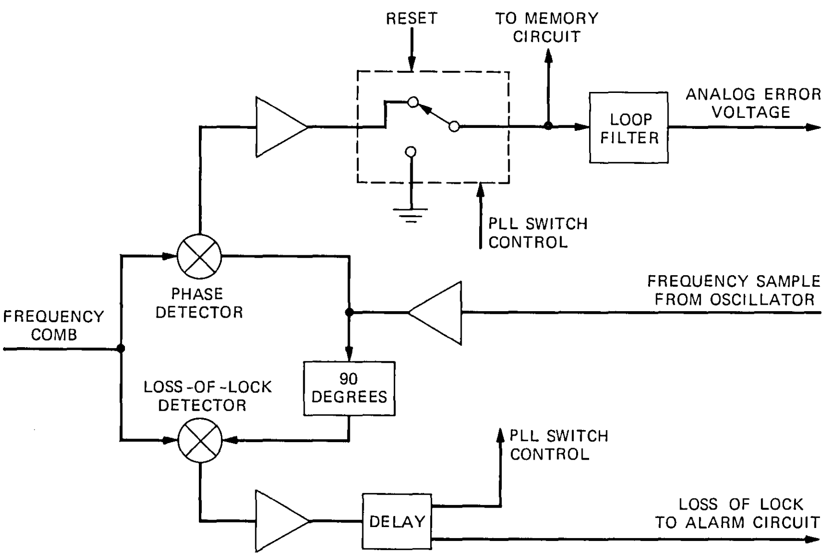

Fig. 9—Frequency control phase-locked loop circuit.

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

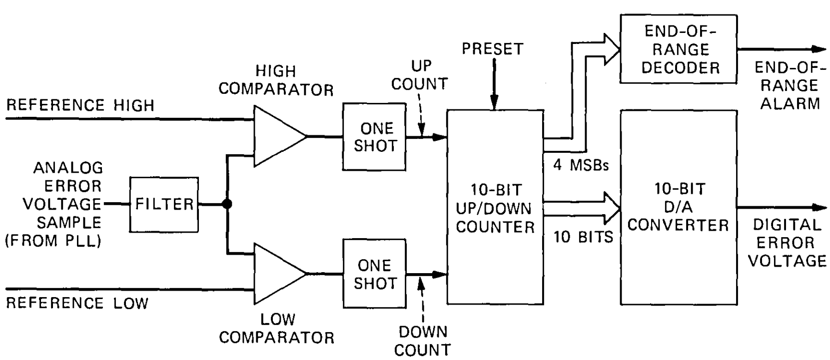

Fig. 10—Frequency control-unit memory circuit.

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.

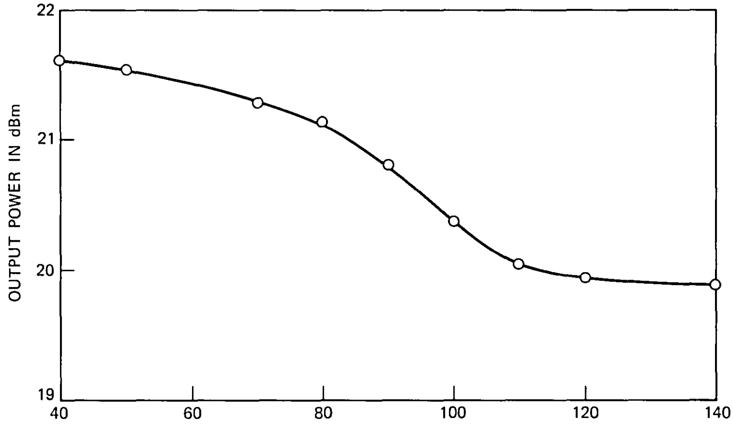

Fig. 11—Output power of the microwave generator as a function of ambient temperature.

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.

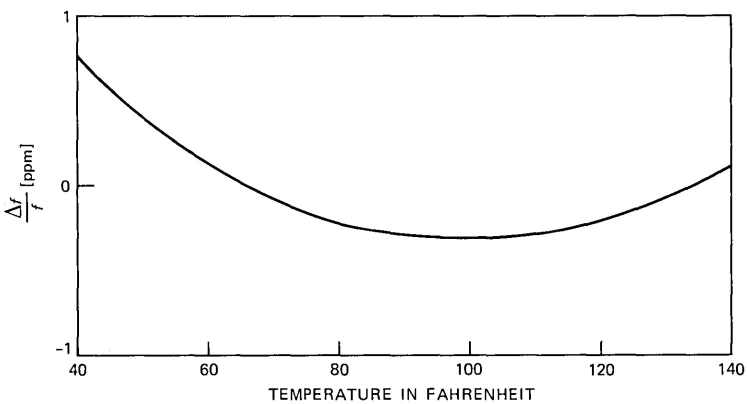

Fig. 12—The measured free-running frequency stability of the microwave generator as a function of the ambient temperature.

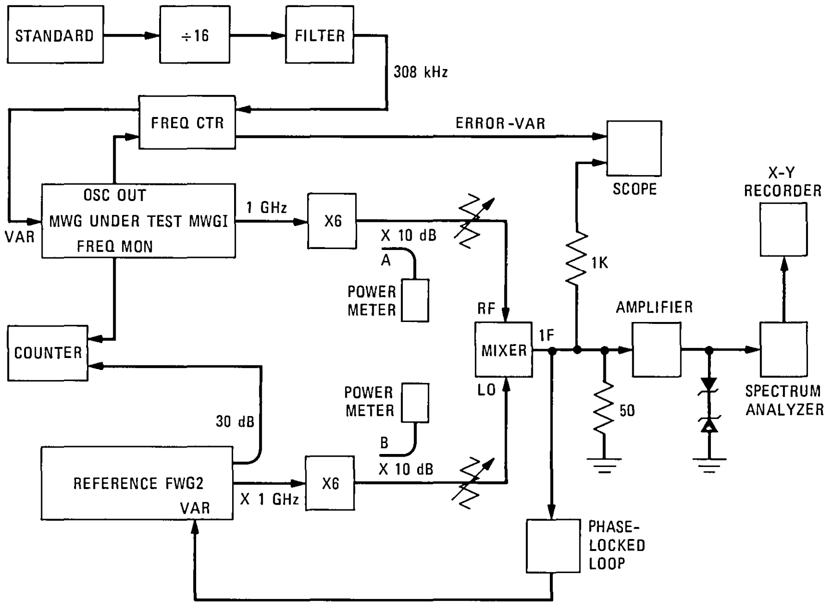

Fig. 13—Block diagram of the test set used to measure noise of the microwave generator.

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.

VII. ACKNOWLEDGMENTS

The development of the microwave carrier supply was the combined accomplishment of many individuals. Among them the authors wish to acknowledge the contributions of T. J. Case, H. Goldstein, and J. R. Scoville.

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The AR6A Single-Sideband Microwave Radio System:

Equalization for Multipath Fading

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Multipath fading can introduce severe amplitude distortion and level changes in a radio channel. These must be dynamically equalized to meet the toll transmission requirements of AR6A. This article describes equalizer circuitry at intermediate frequencies, which continuously senses the level as a function of frequency in the transmission band and dynamically corrects the effects of multipath fading.

I. FADING CHARACTERISTICS

In line-of-sight microwave radio transmission, the broadband radio channels (20 to 30 MHz) can exhibit the phenomena of both selective and nonselective fading. During nonselective fading, the signal power across the channel remains constant with frequency and simply decreases in level. This type of fade can be caused by attenuating effects of the atmosphere or it can be the precursor of selective fading. Some atmospheric conditions can cause propagation over two or more distinct paths, resulting in the reception of multipath components.[1] This event can cause the channel to experience selective fading. During selective fading, not only does the channel show a decrease in received signal power, but the signal level measured as a function of frequency (frequency response) also contains one or more minima.[2] Reference 2

* Bell Laboratories.

contains graphs of measured field data showing typical time-varying frequency responses of a channel during selective fading.

A fade model consisting of four to six multipath components can match most observed channel characteristics.[3] However, a two-path fade model closely approximates a large number of observed channel characteristics.[4] The two-path model has the transfer function:

$$H(f) = 1 - re^{-j2\pi(f-f_0)\tau},$$

where f_0 is the frequency of maximum fade (i.e., minimum signal level) and τ is the delay difference in the two path lengths. The expression $-20 \log (1 - r)$ is the fade depth at frequency f_0.

Computer simulations were used to show that a two-path fade with a fade depth of 20 dB, a time-delay difference of 4 ns, and all possible frequencies of fade maximum (f_0) matches adequately the amplitude characteristics of the worst fades that would have to be equalized to meet system outage objectives for AR6A.*,† An equalizer that could compensate for such a fade to within ±2 dB of the nominal level for all values of the fade center frequency, f_0, will meet system specifications. Some examples of channel characteristics undergoing a two-path, 20-dB, 4-ns fade are shown in Fig. 1.

A good approximation of this selective fade characteristic can be obtained using a power series expansion truncated after the quadratic term: $H(f) = A_0 + A_1(f - f_m) + A_2(f - f_m)^2$, where f_m is the midchannel frequency. The flat term, A_0, can be compensated using an Automatic Gain Control (AGC) amplifier leaving the shaped component alone to be equalized.

From the two-path model of the fading channel, the range of the linear term, A_1, must be ±18.5 dB and the range of the quadratic term, A_2, must be 17 dB. However, large values of the quadratic and linear shaping coefficients are not required simultaneously. This allowed the use of two Bode "bump" networks[5] with center frequencies at the channel ends to jointly realize the linear and quadratic correction terms.[6] It was further determined that a ±10 dB range on each of the Bode "bump" networks is sufficient to meet system requirements on shape equalization. Figures 2 and 3 depict the approximation of linear and quadratic correction functions by Bode "bump" networks. This technique has many advantages with respect to dynamic range and control circuitry, as well as noise figure and intermodulation distortion.

The remaining system equalization objective concerns the rate of change of the channel shape as a selective fade sweeps through the channel. Field data indicate that during deep selective fading, the rate

* Amplitude Modulation Radio at 6 GHz for the initial (A) version.

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

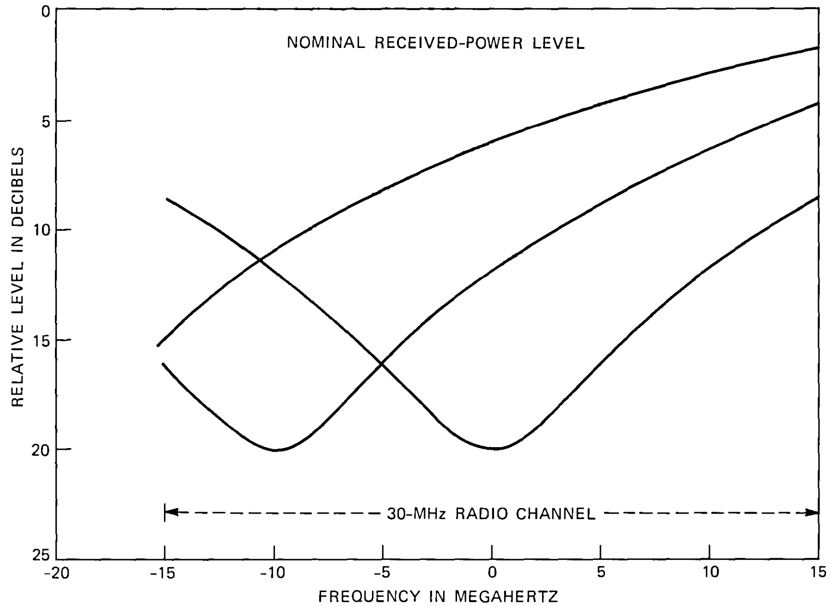

Fig. 1—Typical received-power frequency characteristic for the two-path model of a 20-dB, 4-ns fade in a 30-MHz channel.

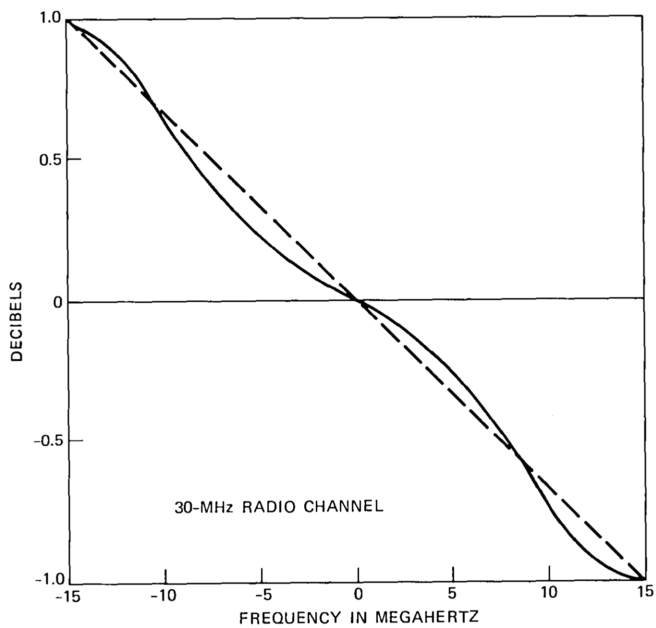

Fig. 2—Summation of two Bode "bump" networks to approximate a linear shape.

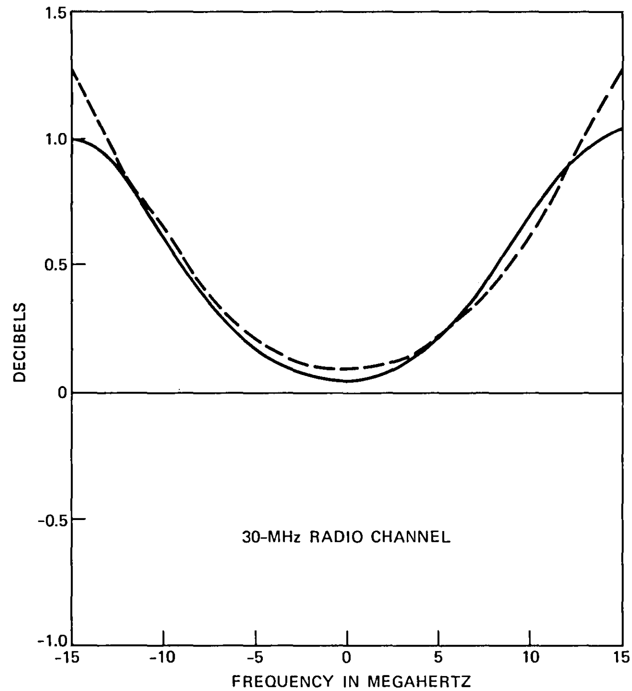

Fig. 3—Summation of two Bode "bump" networks to approximate a quadratic shape.

of change of signal level can be as high as 90 dB per second. Since the most severe fades will be eliminated by space-diversity switching, an objective of adapting to a rate of change of 50 dB per second was established.

II. FUNCTIONAL DESCRIPTION

An error-detecting, zero-forcing technique was chosen to dynamically control the variation of the AGC and shape units. Three pilot tones are transmitted in the radio channel, one near each edge and one near the center. The pilot errors are determined by analog processing of the detected pilot levels. The errors thus determined are fed back to an analog circuit that varies the AGC and equalizer shape coefficients to force the detected errors to zero at the pilot frequencies. The control loop also detects when a space-diversity switch or pilot resupply should be initiated.

Figure 4 is a functional block diagram of the units comprising the dynamic equalizer.

2.1 Gain and equalization control

Gain in the intermediate frequency (IF) channel is controlled by an AGC amplifier that is present at both repeater and main stations. The

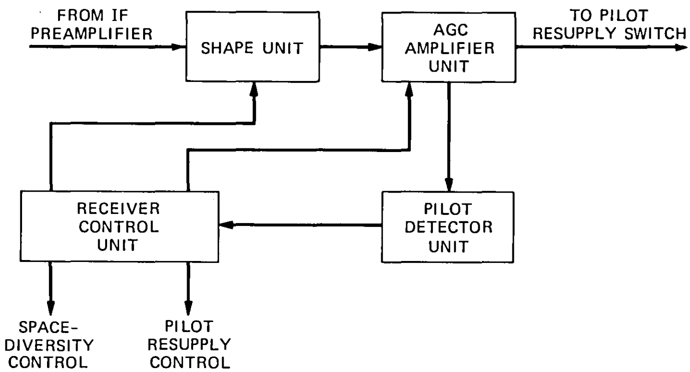

Fig. 4—Block diagram of the AR6A dynamic equalizer.

frequency characteristics of the IF channel are controlled by a shape unit that consists of electronically controlled Bode "bump" networks peaked at the high and low ends of the band. The shape unit is provided only at main stations.

The detector unit senses the levels of the three equalization control pilots and the pilot resupply enable pilot. Processing of the detected pilot levels is done within the receiver control unit. An error voltage for the low (high) bump is derived by subtracting a reference from the difference of the low (high) pilot level and the center pilot. An error voltage for the AGC is formed by subtracting a reference from the average of the three equalization pilot levels. These error voltages will all be zero when the three equalization pilots are at their nominal level. The error voltages are amplified, low-pass filtered, and used as the electronic control for the AGC amplifier and bump shapes.

Nonlinear shaping is provided in the control characteristics of the AGC gain and bump networks such that the gain (in dB) introduced is proportional to the drive voltages. A closed-loop gain of 100 is provided. A 1-Hz cutoff in the loop filter results in a closed loop response time of 10 ms.

The dynamic equalizer has been designed to regulate the pilot levels to within 0.5 dB of their nominal value under conditions of nominal input. For a 20-dB step change in input, the equalizer will settle to within 2 dB of its final value in under 10 ms. Also, a maximum dynamic tracking error of 2 percent of the fade depth in decibels is achieved for fade rates of less than 50 dB per second.

The dynamic equalizer has been designed to achieve a gain flatness of ±0.15 dB at nominal temperatures and ±0.5 dB over the temperature range of 40 to 120 degrees F.

The dynamic equalizer achieves a maximum noise figure of 20.7 dB and third-order intermodulation coefficient for $A + B - C$ products of −72.9 dB for main-station applications.

2.2 Space-diversity and pilot resupply control

The regulation accuracy requirements dictate that the pilot detector realize a precise measure of the nominal pilot level. A low-precision realization of the control loop would suffice were it not for a requirement to precisely determine the crossing of thresholds used to initiate a space-diversity switch.

A space-diversity switch will be initiated if any pilot at the *input* to the dynamic equalizer is more than 36 dB below its nominal level. To avoid the need for separate detectors to monitor this condition, the space-diversity control is derived within the dynamic equalizer circuitry. It turned out that this requirement limited the control circuitry design.

The detector output is an indication of the output pilot value relative to its nominal value. The control voltages to the AGC and bump networks are an indication of the amount of gain supplied between the equalizer input and output. Nonlinear shaping of the detector, AGC, and bump-network control characteristics yields a linear relationship in units of dB per volt; a weighted sum of the detector outputs and the control voltages yields a measure of the input pilot levels. These characteristics must be precisely controlled to provide accurate threshold comparisons.

In addition to the space-diversity control, the equalizer circuitry realizes the control signal that initiates a pilot resupply. This circuitry utilizes a fourth pilot detector, which senses the level of the resupply enable pilot.

III. DETECTOR AND CONTROL CIRCUITS

3.1 Detector

The detector circuit contains four separate pilot detectors interconnected by a tree of hybrid transformers (see Fig. 5). Each pilot detector consists of a hybrid transformer, an IF bandpass filter, an IF amplifier detector, and an operational amplifier gain circuit.

The IF bandpass filter is a two-section, monolithic crystal filter with a 3-dB bandwidth of approximately ±6 kHz and a 60-dB bandwidth of approximately ±50 kHz. The midband loss is approximately 3 dB. Figure 6 shows a typical characteristic of these filters.

The IF amplifier (see Fig. 7) consists of two stages of a series-shunt feedback pair.[7]

The detector portion of the circuit consists of a Schottky diode. An R-C circuit serves as a filter for the detected signal. This circuit acts essentially as a peak detector that derives a dc voltage proportional to the peak of the input IF signal.

The operational amplifier circuit is used as a buffer between the

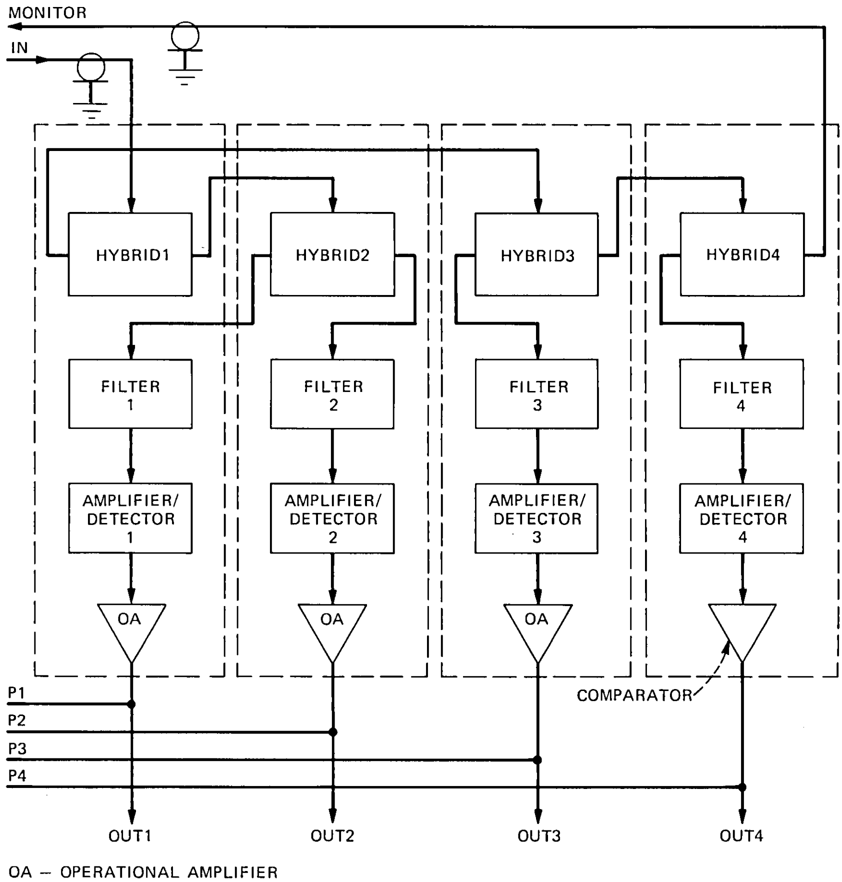

Fig. 5—Pilot detector circuit.

detector and the output of the circuit. This circuit has two adjustments. One is an offset adjustment used to adjust for zero output voltage with a nominal input signal level. The other is a gain adjustment that is set for a given dc output when the input level is 10 dB below nominal. This adjustment is required to compensate for variations in gain at the three pilot frequencies and also for variations in the Schottky diode.

The resupply-enable pilot detector is similar to the three other detectors except that the operational amplifier gain circuit is replaced by a comparator circuit, and the only adjustment is the comparator threshold voltage. A single adjustment is sufficient because only the detection of the presence of the resupply pilot above a certain level is required.

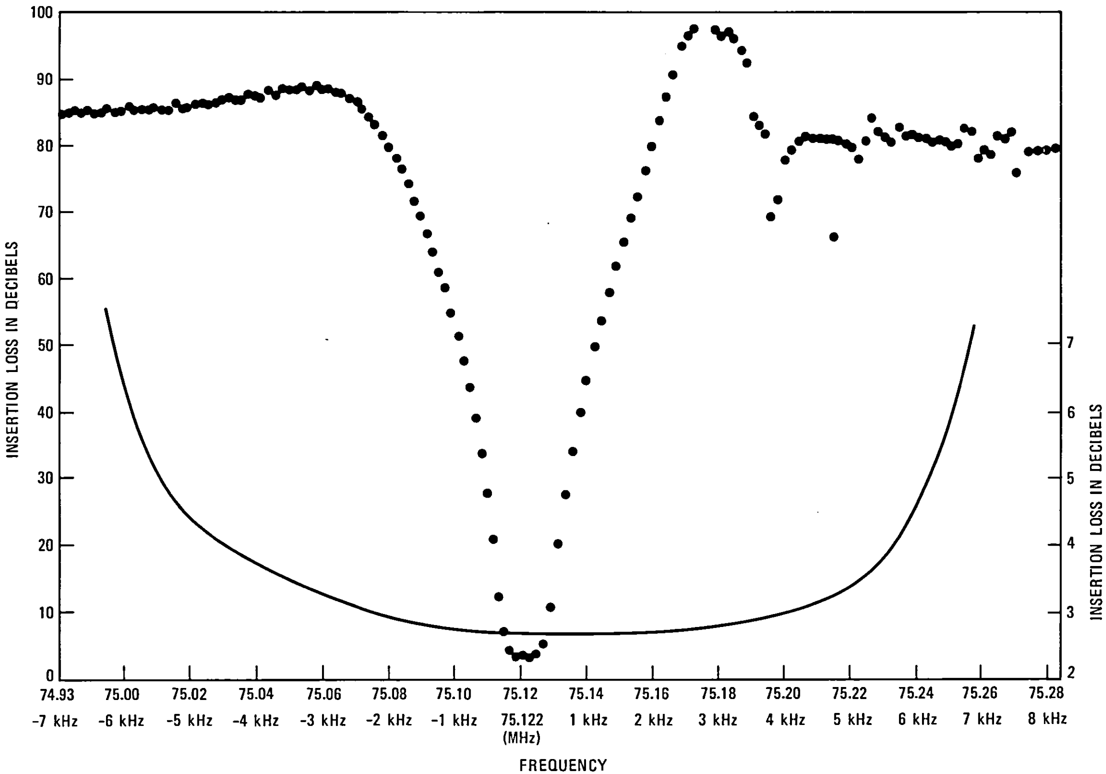

Fig. 6—Typical IF bandpass filter characteristic.

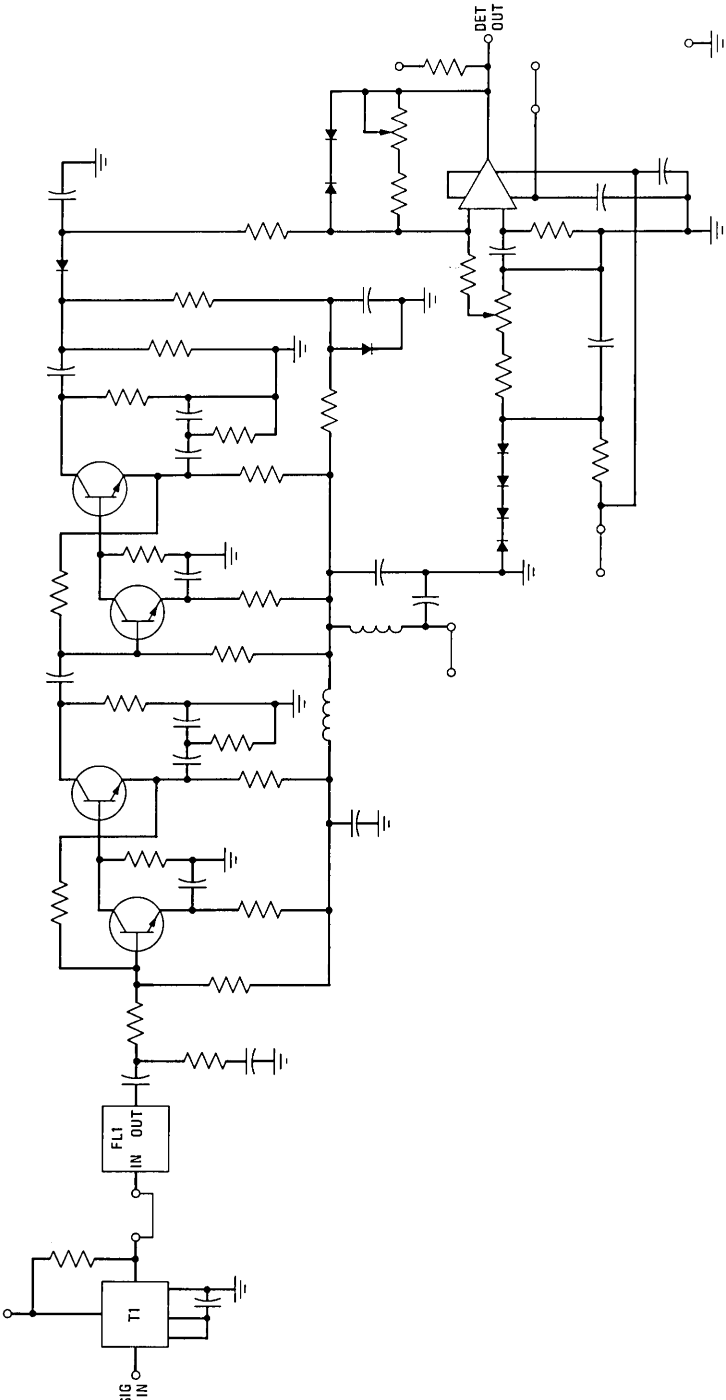

Fig. 7—Pilot detector circuit.

3.2 *Receiver control circuit*

There are two types of receiver control units. The type that only controls the gain of the AGC amplifier is used at repeater stations. The type used only at main stations controls the gain of a shape unit as well as an AGC amplifier.

The receiver control circuit also generates the following initiation signals: (1) a signal to initiate a pilot resupply switch when the receiver is either underpowered or overpowered, and (2) a signal to initiate a space-diversity switch when the level of any one of the three line pilots drops below a certain threshold.

The circuit controls manual and remote receiver gain and remote pilot resupply, and also controls alarm initiating signals when the receiver is in other than its normal mode of operation.

Figure 8 is a block diagram of a main-station receiver control unit. The operation of the circuit is as follows.

To generate the space-diversity switch initiating signal, the inputs

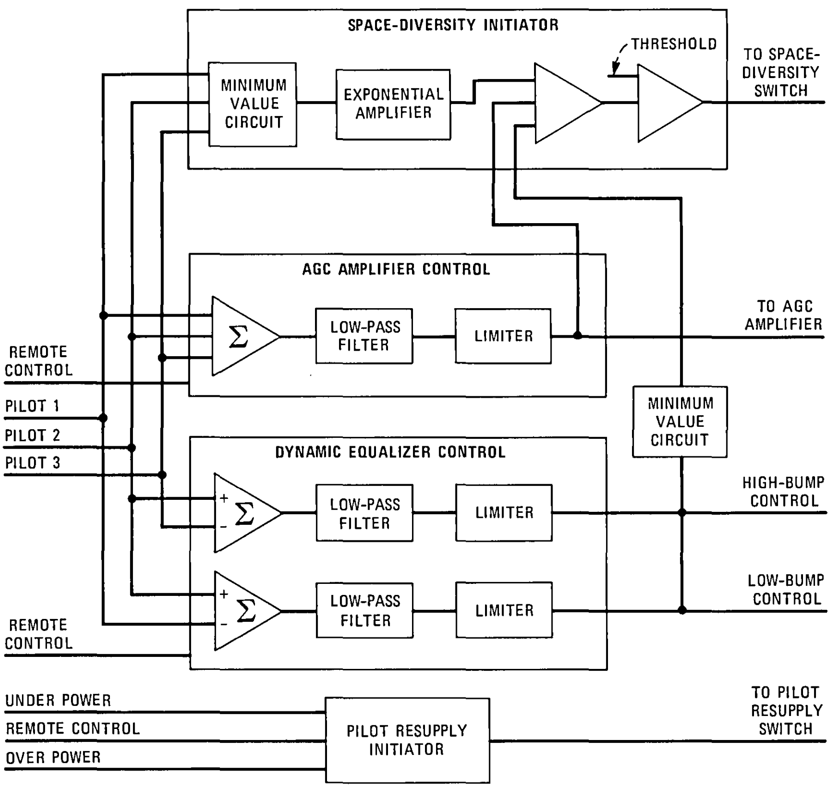

Fig. 8—Main-station receiver control circuit.

from the pilot detector circuits (PLT1–PLT3) are applied to a minimum value circuit. This circuit takes the level representing the lowest level pilot and applies it to an exponential amplifier. The output of this amplifier is a voltage that is linearly proportional to the decibel pilot level. The other input that determines the space-diversity threshold point is the AGC amplifier control voltage. The AGC amplifier is designed so that the gain in decibels is a linear function of the control voltage. The AGC control voltage is determined by the average of the three pilots. The comparator threshold is set to trip when the input voltages are such that the AGC control voltage and the voltage representing the level of the minimum-level pilot indicates that one of the pilots is 36 dB below nominal. The only difference between the repeater-station and the main-station receiver control circuits is that in the main-station circuit the control voltages for the shape unit are also used to determine the space-diversity initiation circuit trip point.

The pilot resupply is initiated in any one of three ways. It will be initiated when the power at the receiver output remains approximately 12 dB above nominal for 100 milliseconds. In this case, the functional input to the receiver control unit is from the resupply switch unit, which senses the total power and produces a control voltage proportional to the average power at the output of the AGC amplifier. The pilot resupply circuit will also be initiated if the average level of the three pilots drops 5 dB below nominal. It releases when the average level returns to less than 4 dB below nominal. This condition is derived from inputs from the detector unit. The pilot resupply can also be initiated remotely via the command and control system.

The AGC amplifier control section takes the voltage proportional to the level of the three pilots, averages them, and low-pass filters the resulting voltage to set the bandwidth of the loop. The output of the low-pass filter is applied to a limiter to limit the maximum gain of the AGC amplifier.

The shape unit control section of the receiver control unit takes the difference between the level of the center pilot and the upper or lower end pilots to control the appropriate bump equalizer in the shape unit. These voltages are also low-pass filtered and limited.

The repeater station receiver control unit does not contain a shape unit control section because a repeater station does not use a shape unit.

The receiver control unit also allows for the AGC amplifier to be set to nominal gain or manually adjusted over its entire gain range. Light-emitting diodes (LEDs) on the faceplate of the unit indicate when the dynamic equalizer is in other than its normal mode of operation.

IV. THE SHAPE UNIT

The shape unit is composed of four Bode-type adjustable "bump" networks separated by amplifiers and controlled by four driver circuits as shown in Fig. 9. Two identical networks provide a bump-type characteristic at 59.8 MHz. The remaining two networks are also identical and peak at 88.5 MHz. Each network section is separated by amplifiers to provide both impedance buffering and gain to offset the flat loss introduced by the networks. High-frequency network sections are alternated with low-frequency sections to minimize interaction between like sections.

Each network section has an associated drive circuit. This provision allows for the independent operation of each network section so that varying p-i-n diode characteristics may be compensated in the driver element of the network section.

4.1 The bump equalizer sections

Each "bump" network section is composed of a series-type, adjustable Bode network as shown in Fig. 10. The adjustable element is a p-i-n diode. The p-i-n diode acts as a resistor that varies in accordance with a bias voltage supplied by the drive circuit. As the resistance value of the p-i-n diode varies, the network shapes the amplitude of the transmission above and below the flat loss level. Measured amplitude shapes are shown in Fig. 11.

4.2 The buffer amplifiers

As shown in Fig. 12, all of the amplifiers provide hybrid-type inputs and outputs. All amplifiers are designed to present 75-ohm input and output impedances. The amplifiers are, however, tuned to operate against the 301-ohm impedance presented by each "bump" network section. Each amplifier provides 7 dB of gain to compensate for the loss introduced by the bump sections.

4.3 The drive circuits

The drive circuits receive the drive voltages from the receiver control unit. As shown in Fig. 9, two voltage follower circuits isolate the drive circuits from the receiver control unit. A schematic description of the drive circuits is shown in Fig. 13.

Each drive circuit is a unity gain dc amplifier. Potentiometer R26 allows for the adjustment of the offset voltage to the input of the amplifier. It is adjusted to provide flat transmission over the 59- to 89-MHz band with a drive voltage of 0V.

Elements R13 through R20, CR2, and CR3 function to make the insertion loss in decibels at the 62.448- and 85.856-MHz pilot frequencies proportional to the drive voltages. The circuit is adjusted so that

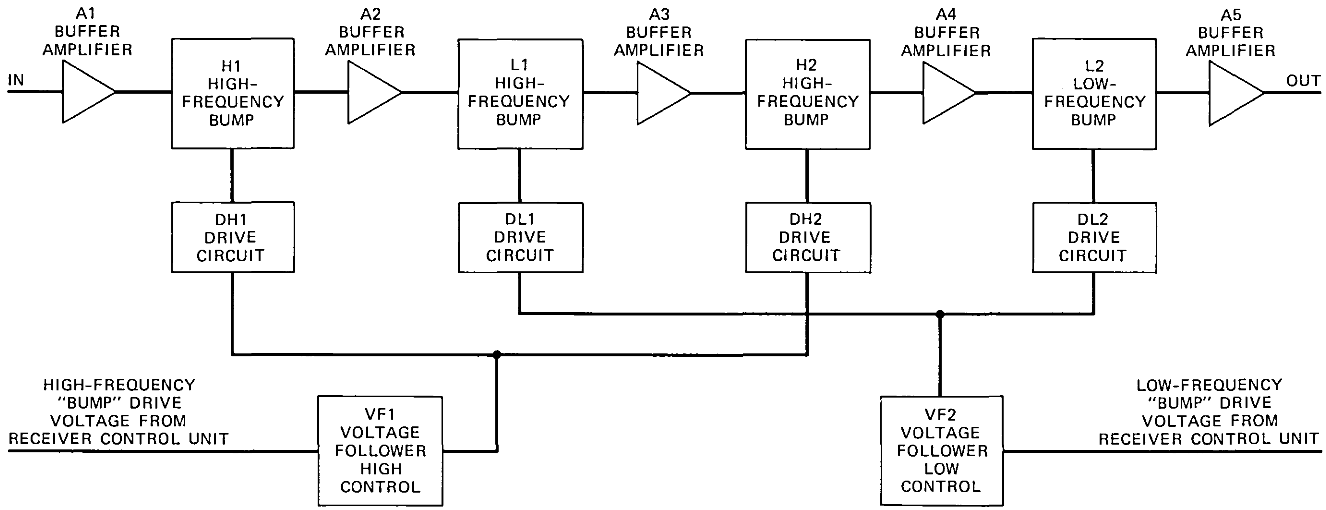

Fig. 9—Block diagram of the dynamic equalizer.

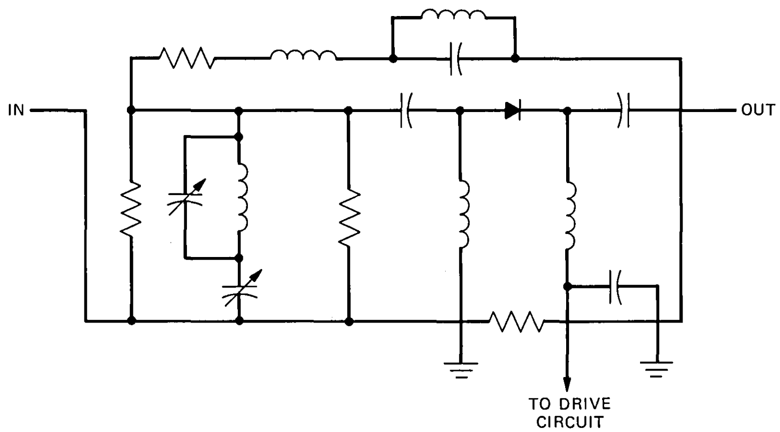

Fig. 10—Bode-type adjustable equalizer.

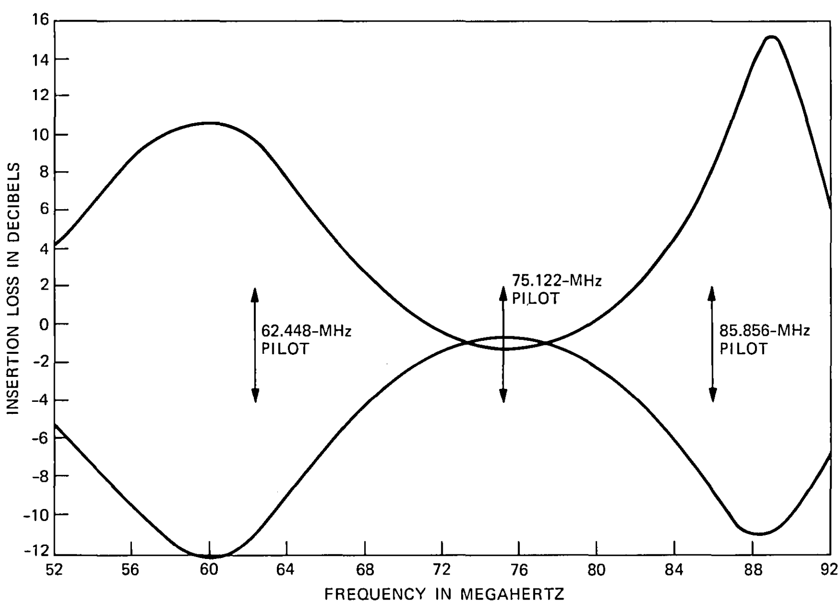

Fig. 11—Typical amplitude shapes generated by the dynamic equalizer for the AR6A System.

a change of 1V in drive voltage results in a 1-dB change in transmission for each network section. This linear relationship is the criterion upon which the space-diversity switching is based.

Diodes CR4 through CR6 act as temperature-sensing elements. Their purpose is to compensate for the temperature versus resistance characteristic of the p-i-n diode. Over the 40- to 140-degrees F temperature range, amplitude distortions in the transmission amount to

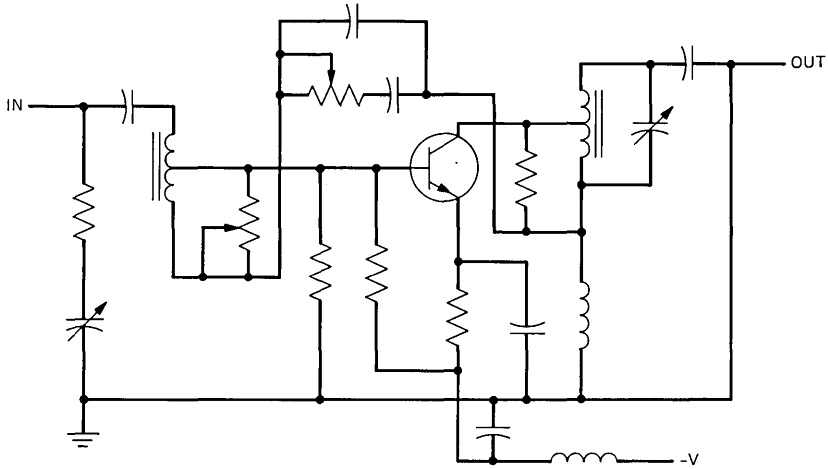

Fig. 12—Buffer amplifier.

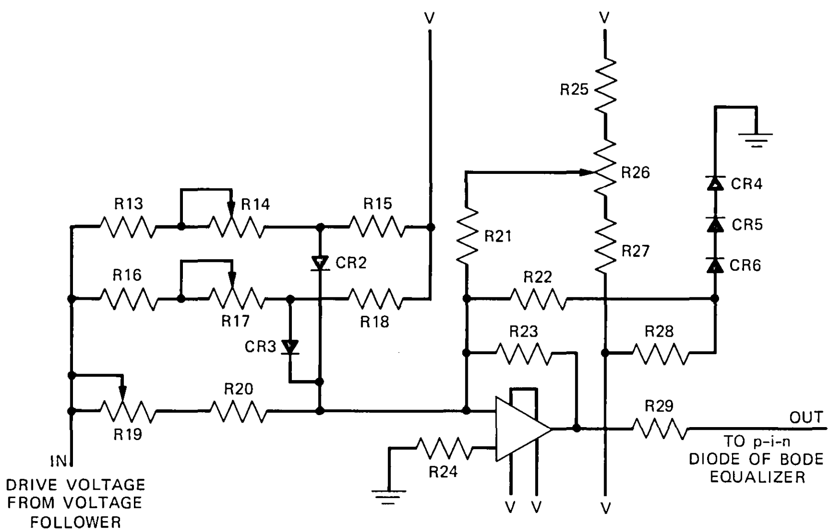

Fig. 13—Drive circuit.

as much as ±1.5 dB between pilot frequencies without compensation circuitry. With compensation these distortions are held to within ±0.25 dB.

V. AUTOMATIC GAIN CONTROL AMPLIFIER

The AGC amplifier normally operates at a nominal gain of 15.7 dB and has a gain range of 61 dB. It is made up of a series of fixed-gain amplifier blocks alternating with variolossers (see Fig. 14). Each

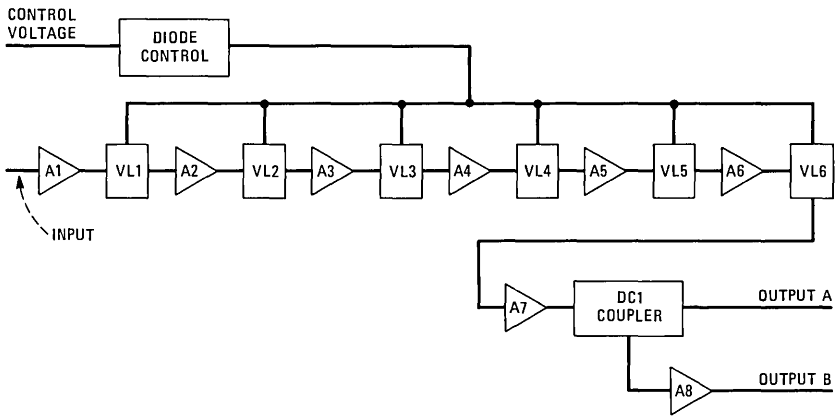

Fig. 14—Block diagram of an AGC amplifier.

variolosser has a dynamic range of approximately 10 dB, so six are necessary to provide the required 61-dB overall control range.

A control circuit is provided to convert the AGC input control voltage to a form that can be used to control the loss of the variolossers. This circuit is basically a voltage-controlled current source that linearizes the input control voltage versus overall AGC gain transfer curve. Another function of the control circuit is to stabilize the temperature of the AGC amplifier.

5.1 Amplifier blocks

Eight fixed-gain amplifiers of the same basic design are required in the AGC amplifier.

Amplifiers A1 through A6 (see Fig. 14) are two-stage feedback amplifiers with a gain of 10.5 dB each. Amplifier A7 has a gain of 16 dB. This gain distribution was chosen to optimize the noise figure and intermodulation performance of the AGC amplifier. Very careful attention had to be paid to the physical design and layout since there is a total distributed gain of almost 80 dB.

The buffer amplifiers in the AGC saturate at a rather high power level due to the heavy biasing required to achieve low intermodulation distortion. Interstage clamping circuitry was added to the AGC to limit the output power to a safe level without affecting normal operation.

Directional coupler DC1 samples a portion of the main path signal and directs this signal to the amplifer A8. Amplifier A8 is a single-stage feedback amplifier with a gain of 8.5 dB. This output provides the signal to the pilot detector unit.

All eight of the amplifier blocks utilize a hybrid transformer impedance-matching configuration (see Fig. 15). This circuit can easily

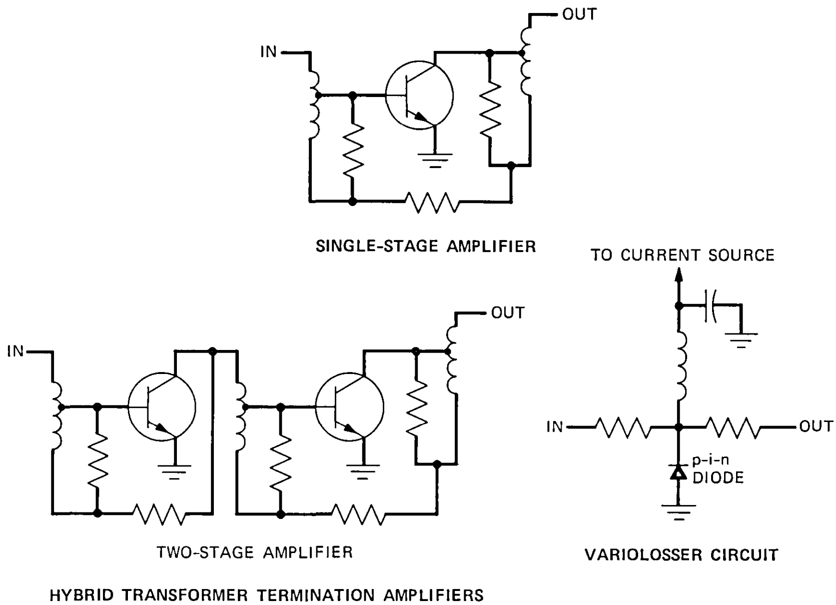

Fig. 15—Functional diagram of the AGC amplifier.

provide better than 30-dB return loss, which is important for the input and output ports of the AGC amplifier and for providing good buffering characteristics to mask the variolosser impedance variation.

The amplifiers in the AGC have been designed for optimum linearity performance to minimize intermodulation distortion. All eight amplifiers use a transistor that has been designed for optimum linearity. To utilize this transistor properly it must be operated at a collector voltage of 13V and an emitter current of 120 mA. Since 15 transistors are required for the AGC amplifier, the result is a relatively high total dc power consumption of 2 amps at −14.7V, or 30W. Consequently, a rather elaborate heat sink and ventilation scheme was required (see Fig. 16). The heat removal problem[8] was complicated by a very tight radio frequency interference requirement on the AGC amplifier.

5.2 Variolossers

The variolosser is essentially a T-pad attenuator with a p-i-n diode as the shunt arm to ground (see Fig. 15). The impedance of the p-i-n diode is varied by changing the bias current through the diode. The p-i-n diode is a type that was developed specifically for this application.

The input AGC control voltage is common to all variolossers (see Fig. 14), but each p-i-n diode has its own voltage-controlled current source. The loss of each variolosser is variable from approximately 3 to 13 dB.

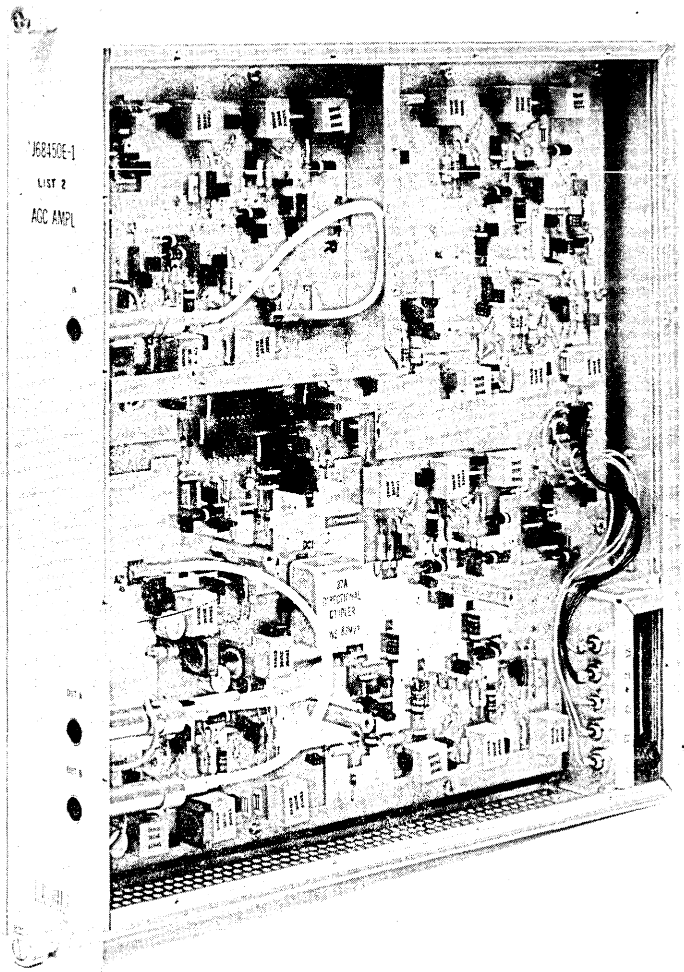

Fig. 16—AGC amplifier.

5.3 Control circuit

The control circuit linearizes the control voltage versus decibel AGC gain curve and compensates for the temperature in the AGC amplifier.

It was necessary to linearize the control function since the p-i-n diode variolosser characteristics are very nonlinear. The AGC control section introduces a compensating distortion in the control function so that the overall transfer function is linear.

The p-i-n diodes in the variolossers are temperature sensitive so the ambient temperature variations must be compensated for. The expected ambient variation is from 40 to 140 degrees F. The voltage across a forward-biased silicon diode is used as a temperature-sensing element. This temperature-dependent voltage is used to modify the variolosser drive voltage in the proper way so that the AGC gain remains constant as the ambient temperature varies.

The decision was made to utilize a two-point "switched" temperature compensation concept rather than attempting the formidable task of tracking the constantly changing p-i-n diodes. With this method two points on the AGC gain curve are perfectly temperature compensated with an acceptable smooth transition in between.

VI. CONCLUSION AND ACKNOWLEDGMENTS

A dynamic equalizer consisting of an AGC amplifier, a shape unit made up of Bode "bump" networks, and control circuitry has been designed to correct for the effects of selective fading in an AR6A channel. The use of Bode "bump" networks is key to the realization of a dynamic equalizer that is low in complexity and yet adequately meets system requirements.

Credit is due to many individuals associated with phases of this development. The authors specifically wish to acknowledge J. M. Kiker, Jr., for his guidance of the AGC amplifier development and P. D. Patel for the physical design of the dynamic equalizer units and IF shelf. The support and encouragement of F. J. Witt throughout the project is also gratefully acknowledged.

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The AR6A Single-Sideband Microwave Radio System:

The Traveling-Wave-Tube Amplifier

By J. F. BALICKI,* E. F. COOK,* R. C. HEIDT,* and V. E. RUTTER*

(Manuscript received February 4, 1983)

This paper describes the amplifier for the AR6A radio transmitter, which consists of a Traveling-Wave Tube (TWT), its magnetic focus system, and driving power supply. The amplification of single-sideband modulated signals requires an unusually low intermodulation noise level and relatively low thermal noise power at the amplifier output. Further, owing to the use of predistortion to achieve the overall repeater intermodulation level objective, the amplifier must perform at high stability. These objectives have been met by choosing suitable TWT design parameters, operating the tube far below its saturated power capability, and devoting close attention to the details of tube construction. The power supply provides well-regulated heater, helix, and collector voltages to control the TWT gain. The supply also regulates the beam current. Current detection circuits protect the tube and power supply from tube internal arcs.

I. GENERAL

The development of a traveling-wave-tube amplifier was undertaken to fulfill all proposed operating requirements for noise, signal intermodulation, gain, safety, and mechanical stability. Prior experience with Western Electric Traveling-Wave Tubes (TWTs)† indicated the

* Bell Laboratories.

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

direction of design. Subsequent application in the AR6A* Radio System has shown that the device, designated the 473A TWT, when operated with the power supply described in this article, meets all detailed repeater requirements. The use of this TWT was discontinued following the decision to terminate the manufacture of all Western Electric traveling-wave tubes. An equivalent TWT obtained from a commercial source and meeting the same requirements is now used in production of the AR6A Radio System.

II. THE TRAVELING-WAVE TUBE

2.1 Requirements on the TWT

The 473A TWT includes the vacuum tube and the focusing magnetic circuit (Fig. 1). This combination must be united at the manufacturing location, so that optimum focus and Radio Frequency (RF) operating parameters are achieved.

The following list of major requirements specified for the TWT includes the operating conditions specified over the frequency band of 5.925 to 6.425 GHz (with full-load output power of 24.8 dBm and gain reduced 2 dB below the optimum value by increasing the helix voltage):

1. RF gain—40 to 48 dB.
2. Gain plus noise figure—69.5 dB maximum.
3. Third-order intermodulation (three-tone test method)[1]—M_{A+B-C} = −90.5 dB maximum, and constant within ±0.4 dB with the power per tone varied from 18 to 30 dBm.
4. Fifth-order intermodulation ($M_{2A+2B-C}$)— from 22- to 30-dBm output power, −180-dB maximum at 18-dBm output power, −160-dB maximum.

The device developed to meet these requirements consists of a helix-type, single-collector, convergent-flow, electron-gun traveling-wave tube, focused by a Periodic Permanent Magnetic (PPM) field.

As observed from the requirements, noise figure, signal intermodulation, and gain were the major operating parameters to be satisfied. Also, implied by the application, long operating life requires low cathode current loading and minimum interception of current from the beam by the helical slow-wave structure. Further, a critical need for this application is mechanical stability; it was verified that minute movement of the helix results in significant shift in intermodulation (IM) performance.

III. TWT COMPONENT DESIGN APPROACH

3.1 Focus circuit and tube envelope

The PPM focus circuit was adapted from an existing Western Electric manufactured circuit used on the 461A TWT. The existing

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

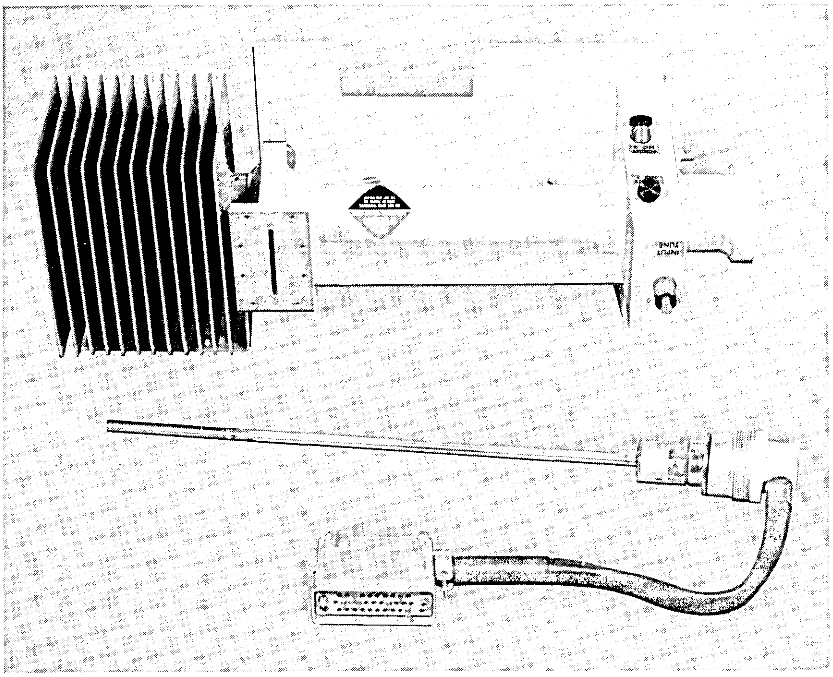

Fig. 1—Picture of AR6A 473A TWT.

glass envelope was also used so that it fit the magnet and pole piece parts. Since the TWT was required to be convection cooled by free air, a finned cooling structure was developed. An SMA coaxial RF input was specified and the coaxial-to-waveguide transition was made an integral part of the circuit.

Of major importance was the introduction of a cross-magnetic field in the collector region. The reason for this is that elastically reflected electrons or electrons emitted by secondary emission from the tube collector resulted in feedback to the input, which has a deleterious effect on intermodulation performance. The smoothing effect, with power variation, upon the third-order intermodulation, with and without the use of a cross field, is shown in Fig. 2.

The focus circuit is assembled from Alnico 8 magnets and soft iron pole pieces on precision mandrels so that no trimming of the magnetic field is necessary for tube focus. A movable pole piece near the input waveguide allows optimum focus to be reached. Cooling fins are coupled to the copper collector of the tube via a thin layer of conducting silicone grease. Adjustable tuning plungers in the waveguide optimize the RF impedance match between the input and output waveguides and helix.

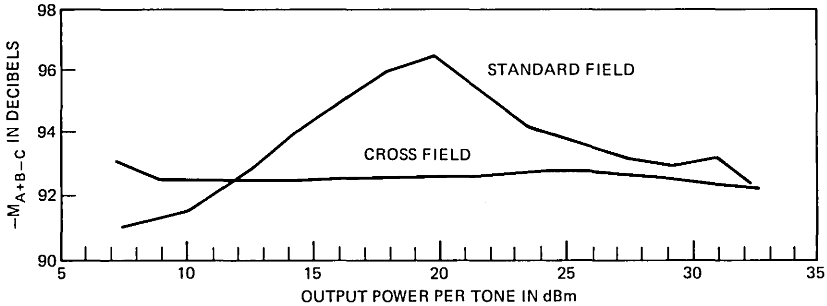

Fig. 2—Effect of crossed-field magnet on M performance.

3.2 Electron gun

Design of the electron gun and its application in the 473A is a departure from previous radio relay tubes because of noise figure requirements. In other Western Electric designs, major guidelines had been avoidance of cathode damage due to bombardment by heavy positive ions and use of moderate cathode loading. To afford ionic protection, previous practice was to design the gun so that the anode operated at higher voltage than the helix, thus, erecting a barrier to backward-flowing positive ions. In addition, the devices were built with highly convergent beam profiles, i.e., with the cathode diameter being much greater than the diameter of the beam projected through the helix. Typically, the emission density in such designs is $0.2A/cm^2$ of cathode area.

The effect on noise figure of both helix-voltage-to-anode-voltage ratio, E_w/E_c, and cathode current density, J_C, is shown in Fig. 3. The upper plot, evaluated at $J_C = 0.2$, shows the noise figures obtained with previous design rules. Clearly, lower noise figure can be obtained with $E_w/E_c > 1$ and with the cathode diameter smaller. A standard oxide cathode coating would be advantageous over other less proven, high-current types if the cathode loading could be reduced to allow its use. The anode voltage was set at 3150, nominal, so that $E_w/E_c = 1.3$. Diminishing benefit would be obtained from a higher ratio, and leakage problems on insulators would also increase. The resulting electron gun is a glass-rod supported triode that emits 70 mA dc from an oxide-coated cathode. The cathode-heater structure is designed to operate at 740 degrees Celsius (observed pyrometric temperature).

The life capability of the cathode has not been entirely confirmed. Based on calculation and extrapolation, greater than 30,000 hours is predicted. Limited facilities allowed five tubes to be run individually over times ranging from 3100 to 8500 hours. One of the five showed significant cathode degradation at 6000 hours; the other four were stable. No short-life problems have been encountered in application.

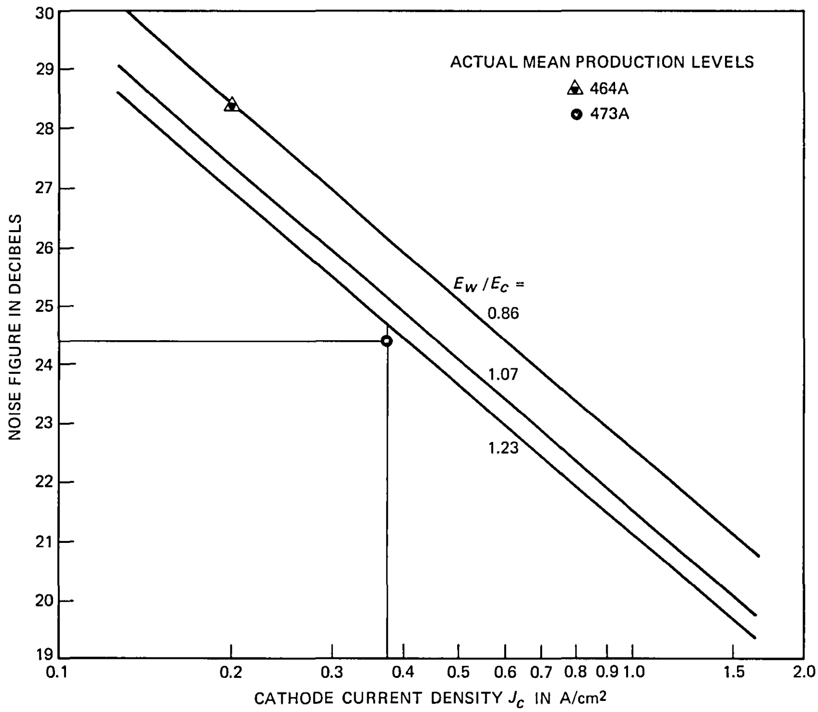

Fig. 3—Effect of cathode loading on noise figure for different voltage profiles.

3.3 Helical slow-wave circuit

Initial tests confirmed the work of Curtice,[2] showing that for lowest intermodulation distortion the helix RF loss should be minimized and the gain parameter C (which is proportional to the 1/3 power of beam-current-to-helix-voltage ratio) should be maximized.

Consideration of the effect of the gain parameter led to design of the helix to operate at 4100 volts. It is noted also that minimum helix loss is desirable for minimizing tube noise figure, although the effect is not as pronounced as it is on intermodulation. Low loss and mechanical rigidity tend to be opposed, because additional support material adds to RF losses.

The helix structure, therefore, utilized beryllium oxide rods. After refining the processing, rods with round cross section had nearly as low loss as those with modified T-shape cross section. The round cross-section rods were used and were bonded to the helix wire with glass frit (glaze). This arrangement was considered the best compromise from the standpoint of intermodulation and noise performance, mechanical rigidity, and heat dissipation.

The helix sever loss (i.e., the loss deliberately added to the midsec-

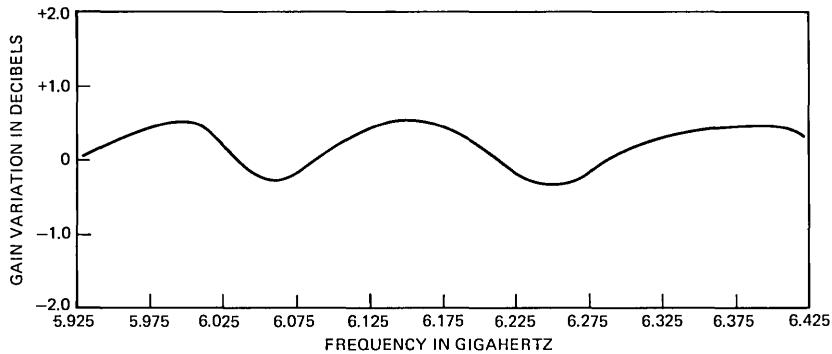

Fig. 4—Typical 473A traveling-wave-tube broadband gain variation.

tion of the helix to prevent self-oscillation and to provide good external RF matches) is a tantalum-aluminum sputtered film with Ta/Al ratio of 70/30 atomic percent.* This loss section is highly stable and not easily altered by bombardment with the 4000-volt electrons of the TWT beam.

The taper applied at the beginning and end of the sever loss is critical to achieving the desired broadband transmission shape. Figure 4 shows the typical broadband gain characteristic of the TWT. The required variation must be less than 1.2 dB peak to peak with no more than four cycles in the 500-MHz band.

3.4 *Collector*

Construction similar to that employed in other Western Electric designs was used. The material is copper, of a length such that the modulated beam is well within the collector before significant beam spread occurs. This point is important from the standpoint of secondary emission. As noted previously, control of secondary electrons and reflected electrons is obtained by a cross-magnetic field over the collector.

3.5 *Processing*

The design included careful attention to parts cleaning, degassing, and handling. Final exhaust was conducted with titanium gettering pumps and seal-off of each vacuum envelope was accomplished only after indicated internal pressure was in the region of 10^{-8} mm Hg.

IV. TWT DEVELOPMENT PROBLEMS

Performance of the TWT was evaluated on a number of models. Problem areas that would likely result in low yield in manufacture were identified and changes were made.

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*This film was originally developed for film resistors.[3]

4.1 *Electron gun interelectrode leakage*

Initially, a gun structure using ceramic rod supports similar to that used in the Western Electric 461A[4] TWT was used. It was soon found that the higher voltages resulted in electrical breakdown across the support rods. A glass rod support structure was designed, leading to much improvement in the leakage problem. It was also necessary to maintain careful parts cleaning and processing. To keep electric fields low, adjacent parts, such as the gas getter and leads, were precisely placed and an initial voltage treatment of the gun was incorporated. The power supply was designed, as later described, to be insensitive to arcs that may occasionally occur. These actions resulted in attaining control over leakage and arcing problems.

4.2 *Low-frequency noise*

The output of early tubes often showed noise at a few hertz and a few tenths of a decibel in amplitude. A rather elaborate computer-based study indicated that the noise was due to oscillation of ions trapped in potential depressions in the electron gun. The study also gave direction to gun design changes required. However, during the period of this study, the gas getter was changed from a nonevaporating type to an evaporating type; the change was actually made from the considerations of lowering cost and using less space within the envelope. It was found that the oscillation problem disappeared, presumably because the new getter reduced the pressure of the offending ion. Though attempts were made, the ion responsible for noise generation was never identified.

V. TWT RESULTS

The gain and third-order intermodulation coefficient were characterized using a special three-tone intermodulation test set.[5] Several hundred tubes (including preproduction versions) were characterized. Typical results are shown in Figs. 5 and 6, which plot the gain and M_{A+B-C} versus power and helix voltage, respectively. These measurements versus power were particularly important because of the requirement for constant third-order performance, which is necessary for the application of the predistorter to the AR6A transmitter.

Sample distributions of three major operating parameters from production—gain, third-order intermodulation, and noise figure plus gain sum—are shown in Figs. 7 through 9. Excellent control within the specification requirements is demonstrated. The remaining important parameter, fifth-order intermodulation, was measured only occasionally and was always well below the specification limits.

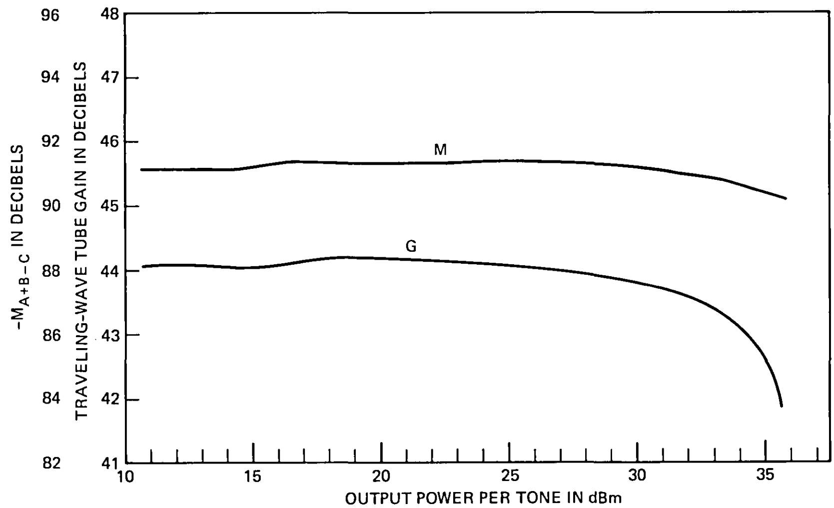


Fig. 5—Typical results of 472A TWT performance study showing third-order intermodulation coefficient and gain versus RF output.

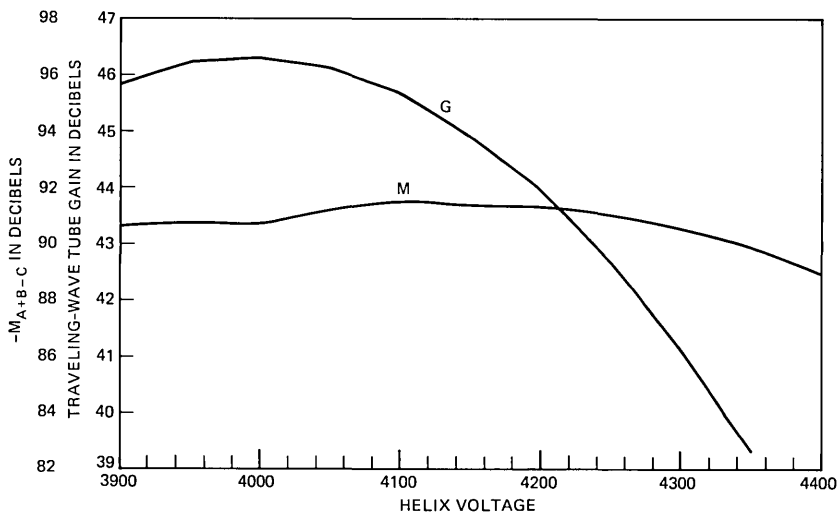


Fig. 6—Typical results of 472A TWT performance study showing third-order intermodulation coefficient and gain versus helix voltage.

VI. REQUIREMENTS ON THE POWER SUPPLY

The requirements for the TWT power supply, with emphasis on those features that directly interact with the TWT, are summarized as follows:

1. Well-regulated heater and helix voltages and beam current are required to control the TWT gain and intermodulation noise.

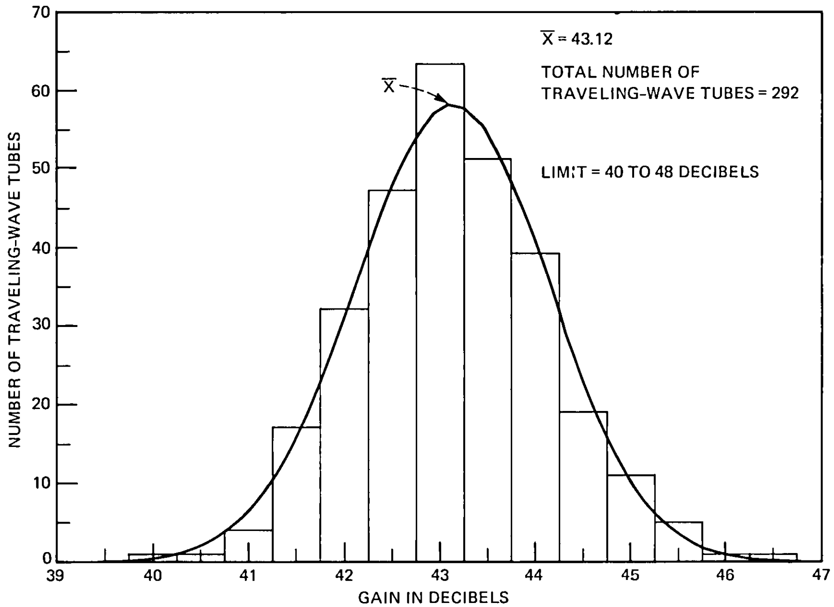


Fig. 7—473A TWT gain.

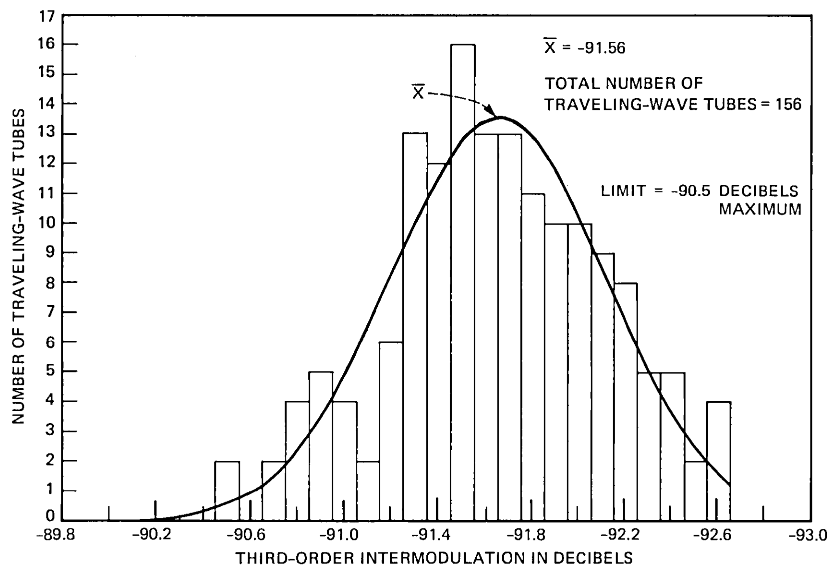


Fig. 8—473A TWT third-order intermodulation ($-M3$).

2. During turnon and turnoff, the electrode voltages must be sequenced so that damaging helix and anode currents do not occur.

3. Current detection circuits must protect the TWT and supply from tube internal arcs.

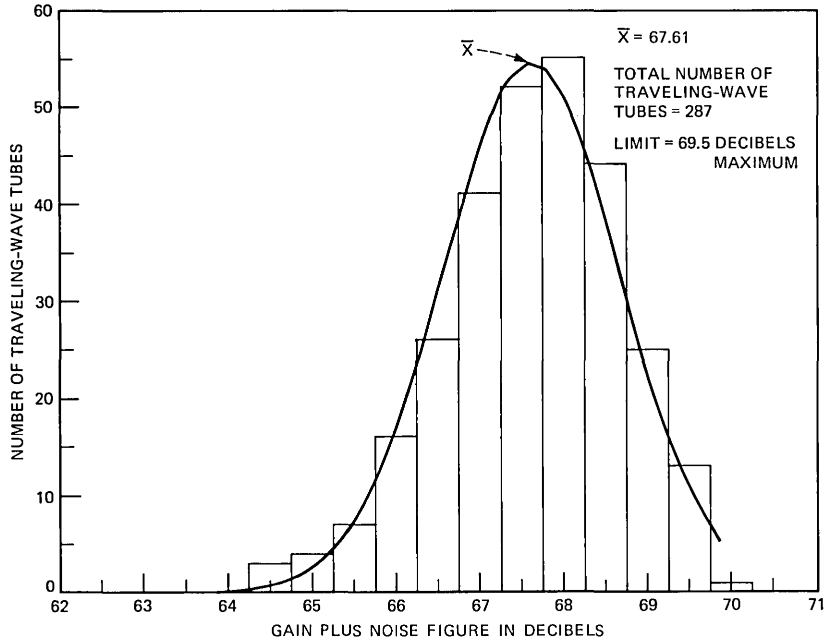


Fig. 9—473A TWT noise figure plus grain sum.

4. Side tones generated by the power supply and TWT combination must be more than 77 dB lower than a single-tone test signal.

VII. POWER ARRANGEMENT AND REGULATION APPROACH

Figure 10 shows a block diagram of the power supply. The power supply and TWT parameters are summarized in Table I. The heater/Beam-Forming Electrode (BFE) dc-to-dc converter supplies 6.3 volts at 1.2 amperes to the TWT heater. The heater voltage is regulated to ±2 percent by controlling the duty cycle of the converter. During initial turnon, the current in the cold heater is limited to about 1.9 amperes by a limit circuit (not shown) within the converter.

The BFE supply is used to turn the TWT on and off electrically. When the TWT is running normally the BFE-to-cathode voltage is less than 1V (nominally 0V). During the time when turnon, turnoff, or internal TWT arcs occur, transistor Q1 is turned off. This causes the BFE potential to be approximately −450V, thus turning the TWT beam off. The BFE switching procedures are discussed in more detail in a later section.

Since the cathode is the common terminal of the TWT amplifier, electrode voltages are, therefore, normally specified with respect to the cathode (even though the collector is grounded). The collector converter supplies +1800V. The 1800V output is controlled by the duty cycle of the converter to achieve ±2 percent regulation.

The anode converter and collector voltages are applied in series combination to provide from 2800 to 3500V to the anode. The beam current is monitored and the output voltage of the anode converter is regulated to control the beam current to 70 mA ±0.5 percent through control of the anode-to-collector voltage. The helix current (I_w) is also monitored. If I_w is less than 1.75 mA, the anode converter regulates the beam current as discussed. Should I_w reach 1.75 mA (corresponding to an unfocused TWT), the anode voltage (E_c) is controlled to hold I_w at 1.75 mA. However, this causes the beam current to fall out of regulation, and when it is 2 percent lower than 70 mA, a trouble indication is generated. This indication occurs when a TWT cathode has become deactivated.

The helix converter and collector converter are applied in a series combination to provide from 4000 to 4550V to the helix. The helix voltage is adjustable using a resistive voltage divider. Since the TWT gain is extremely sensitive to changes in helix voltage, the helix converter includes a series regulator to control the helix voltage to ±0.5 percent. The converter being controlled supplies only the helix-to-ground voltage.

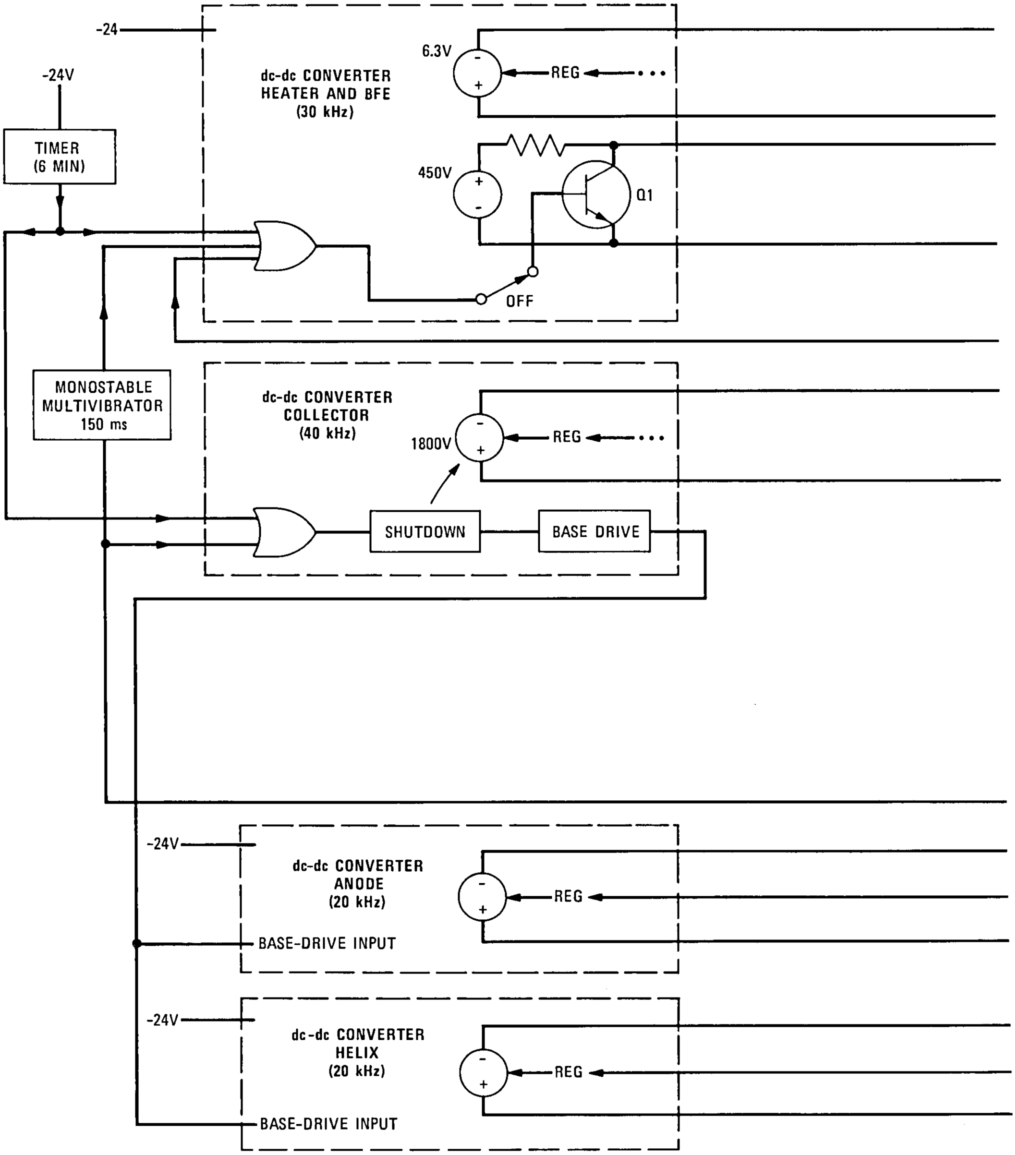
Both the helix converter and anode converter supply only a portion of their respective electrode voltages; the collector supply provides the remaining voltage. This arrangement is used for two reasons. First, the power transformer used in each converter is inexpensive since its secondary must withstand only the voltage it generates itself rather than approximately 1800V plus its self-generated voltage. Second, inexpensive monitoring circuitry may be used since it operates near ground (rather than cathode) potential.

VIII. POWER SUPPLY OPERATION

8.1 Turnon and turnoff

When −24V is connected to the input of the power supply through an external switch, the heater/BFE converter and the timer start immediately. The timer supplies a shutdown signal to the collector converter. Since the helix and anode converters depend on base-drive signals for their inverters from the collector converter, they are also shut down. The timer continues to supply a shutdown signal for a nominal 6 (minimum of 5) minutes to allow the TWT to warm up. During this preheat period the timer also supplies a signal that causes −450V to be applied to the BFE during warm-up to inhibit spurious emissions from the cathode.

At the end of the 6-minute warm-up period, the shutdown signal to


-24
-24V
TIMER
(6 MIN)
dc-dc CONVERTER
HEATER AND BFE
(30 kHz)
6.3V
REG
450V
Q1
OFF
MONOSTABLE
MULTIVIBRATOR
150 ms
dc-dc CONVERTER
COLLECTOR
(40 kHz)
1800V
REG
SHUTDOWN
BASE DRIVE
-24V
dc-dc CONVERTER
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REG
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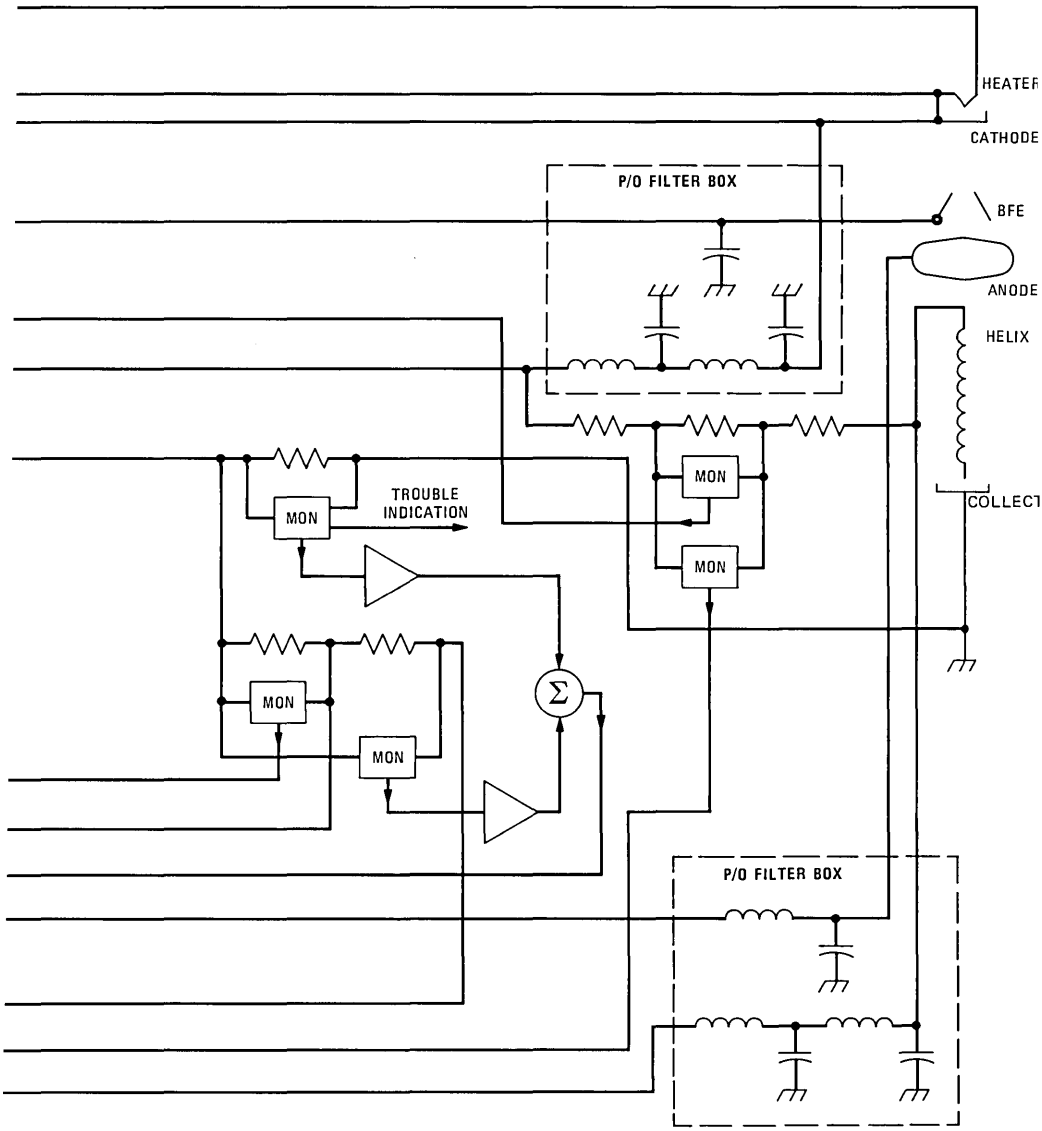

Fig. 10—Block diagram of the TWT power supply.

Table I—Parameters of TWT power supply requirements

Heater current	1.2 amps maximum
Heater voltage	6.3 volts ±2 percent
BFE voltage	0
Anode current	0.5 mA maximum
Anode voltage	As required for collector current
Helix current	1.8 mA maximum
Helix voltage	Adjusted for 5 percent above synchronous operation; regulated to ±0.5 percent after adjustment
Collector current	70 mA
Collector voltage	1800 volts ±2 percent

the collector converter is removed. The collector, helix, and anode converters then start. The time constants of each converter are designed so that these electrode voltages reach their respective final values in the order named above. The BFE voltage is not switched to zero until the collector converter has started, and the helix voltage has reached 4000V. This arrangement prevents excessive helix current while the TWT is turning on.

The external −24V switch is used for normal turnoff. On removal of −24V, the electrode voltages decay, the timer resets, and the BFE supply loses its continuous source of energy. As soon as the helix voltage reaches 4000V, the remaining BFE voltage is applied. There is still enough stored energy in the BFE supply's internal capacitor to inhibit the beam until the other electrode voltages fall safely below levels that would otherwise cause damaging values of helix and anode currents.

8.2 Arc detection

On occasion a TWT arcs. Any combination of electrodes might be involved in the arc, but the direct or indirect result is almost always a large value of helix and/or anode current. When the helix plus anode current exceeds 10 mA, a shutdown signal is applied to the collector converter. Of course the helix and anode converters also shut down, since they depend on the collector converter for base drive signals. Decay of these voltages then causes the arch to extinguish. At this time the collector converter is again turned on and the tube voltages rise toward their respective normal values.

During the arc and subsequent shutdown, the RF output of the TWT is less than normal. When the TWT is turning back on, the process is controlled so that the AR6A protection switching system can operate properly. The RF output is held off by applying −450V to the BFE for at least 150 milliseconds, thus allowing AR6A Radio System transients, which may occur during the arcing process, to subside completely before reapplying RF signal to the system.

The arc detection arrangement described above not only protects the TWT by turning it off for the prescribed time period, but also has proved to be successful in protecting the supply. One particularly arc-prone TWT was life tested for 14 days. A total of 110 arcs occurred, but no supply damage resulted. The TWT parameters gradually fell out of specified limits, but the TWT performance did not change significantly as the result of any one arc.

8.3 Tones

The power-supply ripple requirements are not directly specified. Instead, they are implied by the requirement that the TWT sidetones must be more than 77 dB below the test tone.

Output filters are used in each converter. Additional Inductance Capacitance (LC) sections are mounted in a shielded steel box. A shielded cable connects the output of the filter box to the TWT.

The dominant sidetones occur at 20 kHz, the frequency of the helix and anode converters, and at 30 kHz, the switching frequency of the heater/BFE converter. These sidetones are caused by interwinding capacitance current that is generated in the power transformer in the helix, anode, and BFE converters. These currents are bypassed where appropriate, but they are still the dominant tone sources. In addition, special filtering is provided for transistor Q1 to prevent it from being turned off at a 30-kHz rate by interwinding capacitance current.

IX. AMPLIFIER OPERATION

The TWT and power supply are electrically connected by a shielded cable. At the TWT end, the cable is potted into the end of the tube. The cable connector mates to the power supply connector, which is accessible through a slot cut into the top cover of the power supply. A safety arrangement prevents power-supply operation when the TWT connector is disconnected.

The RF signal connections to the TWT are via an SMA coaxial connector at the input and a reduced height WR159 waveguide at the output. Adjustable shorts are provided for RF matching at the input and output.

Meters on the front panel of the power supply are supplied to measure the helix and collector currents. On the front panel, a helix-voltage adjustment (labeled RF gain) is provided. This adjustment is made during alignment of the amplifier so that the TWT operates at a helix voltage higher than the synchronous value, thus providing gain that is approximately 2 dB lower than that achievable at the synchronous helix-voltage value.

All amplifiers were tested in a bay environment. The gain-versus-frequency characteristic (transmission characteristic) of the amplifier embedded in the AR6A transmitter was measured for each amplifier

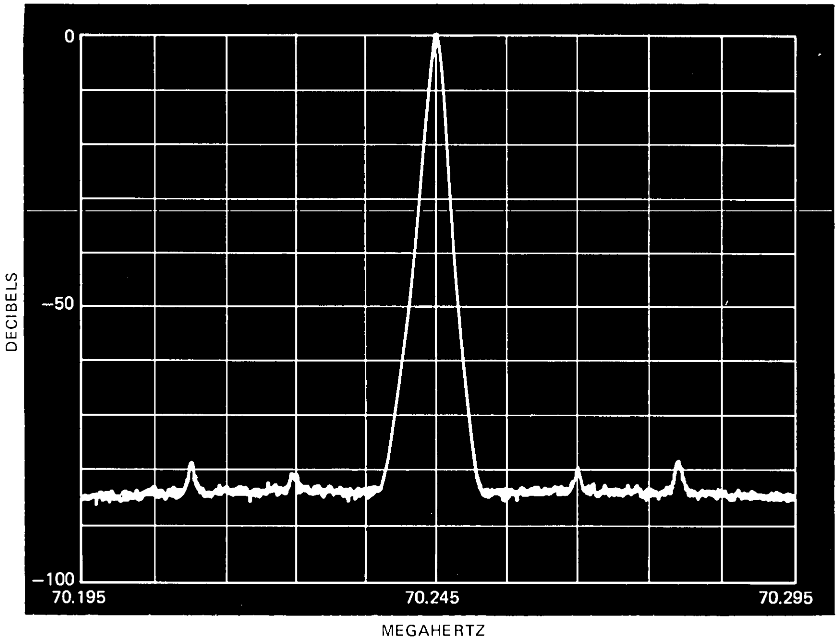

Fig. 11—Test tone with spurious modulation.

shipped for field application. The results of these tests were interpreted by subtracting out frequency characteristics attributable to other transmitter components. Such analysis shows that the average transmission characteristic for 100 amplifiers (with samples at every channel) has a bow of approximately 0.1 dB across the 30-MHz channel. (Note that any residual slope of the transmitter is equalized.) This is approximately what one would expect, considering the gain variation shown in Fig. 4.

Also, a test was made to check for a spurious modulation caused by very low ripple voltages at frequencies that are multiples of the 30- and 20-kHz switching rates of the power supply. A test tone was applied to the amplifier and sidetones were measured. The sidetones were required to be at least 77 dB below the level of the test tone. A graph extracted from a spectrum analyzer photograph shows a typical result (Fig. 11).

X. ACKNOWLEDGMENTS

J. R. Meszar and C. A. Waldron were highly instrumental in the development of the power supply and TWT, respectively.

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The AR6A Single-Sideband Microwave Radio System:

Predistortion for the Traveling-Wave-Tube Amplifier

By R. P. HECKEN,* R. C. HEIDT,* and D. E. SANFORD*

(Manuscript received February 7, 1983)

To satisfy the single-sideband amplitude-modulated system noise requirements, the transmitter must be very linear. Predistortion reduces intermodulation distortion of the transmitter to acceptable levels. This paper deals with the design considerations, the model of the transmitter used to analyze third-order intermodulation products, and the methods applied to align the predistorter-transmitter combination for effective intermodulation reduction.

I. INTRODUCTION

The frequency stability required for single-sideband amplitude-modulated (SSB-AM)† system operation results in the potential for in-phase addition of intermodulation noise from repeater to repeater. Traveling-Wave Tubes (TWTs) offer the best linearity (and, consequently, the lowest intermodulation distortion) available for currently existing 6-GHz power amplifiers. However, linearity must be improved to satisfy system noise requirements.

The concept of predistortion to reduce intermodulation distortion

* Bell Laboratories.

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

had originally been explored extensively at 4 GHz[1] and was found considerably less complex than feedforward.[2] In particular, predistortion at intermediate frequencies appeared attractive because of its simplicity and economy. Operating at intermediate frequency (IF), the predistorter is placed ahead of the up converter and is, therefore, able to cancel intermodulation products generated in the converter and the traveling-wave tube. Clearly, the latter is the major contributor to nonlinear distortions and will be treated accordingly.

Extensive measurements have shown that TWTs, if operated at low radio frequency (RF) power levels, i.e., in a domain of weak nonlinearity, behave essentially as first- and third-order devices and produce odd-order intermodulation products within the narrow frequency band of interest. In the AR6A* System the predistorter is designed to cancel distortions in this power region, and it is important to distinguish this mode of operation from other methods where predistortion is analyzed in a region of strong nonlinearity.[3,4]

In the following section we will proceed with a simple analysis of nonlinearities of the transmitter, which is based on a model that proved to be very accurate within the useful power range of the AR6A System. Section III will deal with practical design considerations, and in Section IV we will conclude with the results.

II. NARROWBAND DISTORTION ANALYSIS

2.1 Nonlinear model of the traveling-wave tube

Measurements on a large number of traveling-wave tubes having a saturated power-handling capability from 5 to 20W showed that even at very low RF power levels (<1W), the simple memory-free polynomial model describing gain nonlinearity of the TWT is inadequate. Two additional effects must be included in the analysis to achieve substantial intermodulation reduction of 20 dB or more: the amplitude and phase distortions due to reactive circuit elements and the amplitude modulated to phase modulated (AM-PM) conversion effect. A model of the TWT that was found very useful is shown in Fig. 1. This model characterizes a special class of nonlinear systems in which a memory-free, nonlinear two-port network is embedded between two linear passive two-port networks having the impulse responses $h_A(t)$ and $h_B(t)$, respectively. For narrowband systems the nonlinear two-port is well modeled by a finite odd-order power series in two variables with constant coefficients[5,6]

$$w = b_1'u + b_3'u^3 + b_5'u^5 + \cdots + b_1''\dot{u} + b_3''\dot{u}^3 + b_5''\dot{u}^5 + \cdots, \quad (1)$$

*Amplitude Modulation Radio at 6 GHz for the initial (A) version.

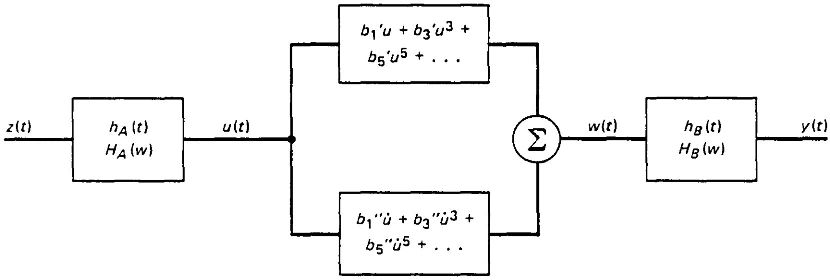

Fig. 1—Approximate model of a traveling-wave-tube amplifier with weak nonlinearities.

where $\dot{u} = du/dt$. The partial series in powers of $\dot{u}$ accounts for AM-PM conversion. An important aspect of this TWT model is the fact that the two interfaces between the three two-ports are, in general, not accessible.

For our analysis, we will limit the power series to first- and third-order terms. Applying the Fourier transformation to eq. (1), we obtain the complex nonlinear input-output relation

$$W = B_1U + B_3U^{3^*} + \cdots, \tag{2}$$

with the notation U^{3^*} indicating the repeated convolution U^*U^*U. The complex coefficients are given by

$$B_1 = b_1' + j\omega_c b_1'', \tag{3a}$$

and

$$B_3 = b_3' - j\omega_c^3 b_3''. \tag{3b}$$

The narrowband assumption manifests itself in eq. (3), in that the frequency variable, ω, has been replaced by the constant channel frequency, ω_c, while performing the convolutions. With the input-output relations for the nonlinear portion of the TWT established, we can write the overall transfer function for the TWT model. Let Z and Y be the Fourier transforms of the input and output signals; then

$$Y = H_BB_1H_ZZ + H_BB_3(H_AZ)^{3^*} + \cdots, \tag{4}$$

where H_A and H_B are the transfer functions of the input and output networks, respectively. Before we proceed with the distortion analysis of the transmitter, we need first to discuss some properties of the up converter.

2.2 The Schottky modulator

Comparative measurements on Varactor Diode Converters (VDC) and Schottky Diode Converters (SDC) have shown the superiority of the latter when considering its nonlinear distortion characteristics versus RF output power. These measurements revealed that the SDC, when pumped with high local oscillator power, behaves like a third-order device over a wide range of RF signal levels. This property is indeed useful because it allows the predistorter, tailored to compensate for third-order distortions, to improve the converter distortions as well. Moreover, because of its conversion loss, in contrast to the conversion gain of a VDC, a higher level of distortions can be tolerated in an SDC. This is so because, for a given performance, the TWT essentially limits the RF output power. To illustrate this, we compute the power of a three-tone, third-order product generated in the converter as it appears at the output of the TWT. If we refer to Fig. 2 this power is

$$P_{A+B-C} = M_{A+B-C} + 3P_A + G \qquad \text{(dBm)}, \tag{5}$$

where M_{A+B-C} is the particular intermodulation coefficient[7] in decibels of the converter, P_A is the power in dBm of each of the equal-level, three-tones at the output of the converter, and G is the overall gain of the bandpass filter (BPF) and TWT. Since the single-tone power at the tube output is

$$P_{AO} = P_A + G \qquad \text{(dBm)}, \tag{6}$$

we find that

$$P_{A+B-C} = M_{A+B-C} - 2G + 3P_{AO} \qquad \text{(dBm)}. \tag{7}$$

Thus, for a specified performance (P_{AO} fixed) the device distortion M_{A+B-C} can be reduced by 2 dB for each decibel increase in TWT gain. Since the difference in conversion loss and gain can be as high as 12 dB, the Schottky diode converter may have an M_{A+B-C} that could be 24 dB worse than that of a varactor diode converter, yet may still yield the same distortion performance. Typically, the SDC has an M_{A+B-C} of about −25 dB compared to −40 dB of the VDC, which clearly shows the disadvantage of the varactor diode converter.

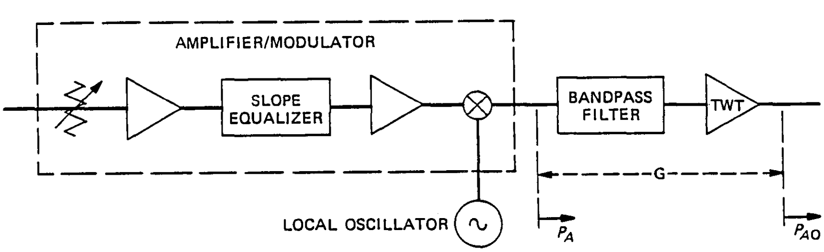

Fig. 2—Simplified block diagram of the AR6A amplifier/modulator with bandpass filter and traveling-wave tube.

It is useful to compare the distortion power generated by the converter to that of the TWT. At a total signal output power of +27 dBm for three equal-level tones, the single-tone power P_{AO} is approximately +22 dBm. With $G = 45$ dB and $M_{A+B-C} = -25$ dB from eq. (7), $P_{A+B-C} = -49$ dBm. A TWT having an $M_{A+B-C} = -91$ dB generates intermodulation power of the same product at a level of -25 dBm. Obviously, one can ordinarily neglect the contribution of the converter since it increases the total intermodulation power by only 0.53 dB. However, if the predistorter can reduce distortions by more than 20 dB, the converter will contribute significantly to the resulting intermodulation noise unless the predistorter also reduces its nonlinearities. In spite of this, to keep our discussion simple we will not include the converter-generated distortions in our analysis.

2.3 Generalized nonlinear transmitter model

The mathematical treatment of the nonlinearities of the complete transmitter is complicated, even when we neglect the converter nonlinearities, because of the operation in two frequency domains. As is shown in the appendix, the frequency translation from IF to RF requires each sideband to be considered separately. We will point out the differences where appropriate and develop the analysis for the upper sideband only. To arrive at a tractable model of the complete transmitter, we actually include the converter, bandpass filter, etc., in the TWT model as described in the appendix. Frequency characteristics at RF are then referred to the IF band, allowing the predistorter to be directly connected to this TWT model. The cascade of predistorter with IF amplifiers and the converter with bandpass filter, TWT, etc., is depicted in Fig. 3, where we used a nonlinear model of the predistorter that is similar to that of the TWT. Thus, we assume that the predistorter has the transfer function

$$Z = G_B A_1 G_A X + G_B A_3 (G_A X)^{3^*}, \tag{8}$$

which leads to the following overall transfer function for the upper sideband (USB):

$$Y_{\mathrm{USB}} = H_B B_1 H_A G_B A_1 G_A X + H_B B_1 H_A G_B A_3 (G_A X)^{3^*} + H_B B_3 (H_A G_B A_1 G_A X)^{3^*} + \cdots . \tag{9}$$

In eq. (9) all terms requiring higher than third-order convolutions have been neglected. Two observations can be made immediately: the distortion terms depend directly or indirectly on the product $G_A X$ as part of the convolution process. Thus, the frequency characteristic of the input network of the predistorter will have no effect on the minimization of these terms. Also, H_B is a factor in both distortion

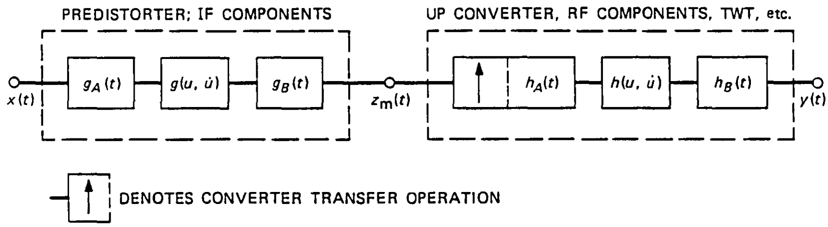

Fig. 3—A generalized model of the AR6A transmitter.

terms, indicating that the frequency characteristic of the TWT output network also has no effect on the intermodulation cancellation. For the purpose of this analysis we can, therefore, set $G_A = 1$ and $H_B = 1$. With these simplifications the third-order distortion term becomes, in the case of the upper sideband,

$$D_3 = B_1 H_A G_B A_3 (X)^{3^*} + B_3 (H_A G_B A_1 X)^{3^*} \qquad \text{(USB).} \tag{10}$$

It can be shown that in order for D_3 to at least vanish over the frequency band where X exists, it is necessary for the product $H_A G_B A_1$ to have constant magnitude and constant delay as function of frequency within that band, i.e., the condition must be met

$$H_A G_B A_1 = C e^{-j\omega\tau}, \tag{11}$$

where

$$C = |H_A G_B A_1| \qquad \text{and} \qquad \omega\tau = \measuredangle(H_A G_B A_1).$$

This yields for D_3

$$D_3 = \left(B_1 \cdot \frac{A_3}{A_1} C e^{-j\omega\tau} + B_3 C^3 e^{-j\omega\tau} \right) X^{3^*}.$$

Hence, frequency-independent cancellation of the third-order distortions at the upper sideband is possible if, in addition to eq. (11), the following condition is fulfilled:

$$\frac{A_3}{A_1} = -\frac{B_3}{B_1} \cdot |H_A G_B A_1|^2 \qquad \text{(USB).} \tag{12a}$$

The corresponding condition for the lower sideband (LSB) is

$$\frac{A_3}{A_1} = -\frac{\check{B}_3}{\check{B}_1} \cdot |H_A G_B A_1|^2 \qquad \text{(LSB),} \tag{12b}$$

where $\check{B}$ is the complex conjugate of B. Thus, the required complex ratios A_3/A_1 for the LSB and USB are conjugate to each other. The fundamental relationships for wideband distortion cancellation are described in eqs. (11) and (12). Therefore, design as well as alignment procedures are essentially based on these criteria. Obviously, eq. (11) requires amplitude and delay equalization for all components in the signal path between the nonlinear element of the predistorter and the nonlinear region of the TWT, unless these components have inherently flat response and constant delay. In the AR6A transmitter, however, a certain amount of slope equalization was required to achieve intermodulation improvement by more than 30 dB. This equalizer is integrated into the modulator/amplifier unit.

From eqs. (12a) and (12b) we derive the required magnitude and phase of the distortions to be generated in the predistorter. The magnitude is practically independent of the sideband used since the TWT generates approximately the same intermodulation level at both frequencies. Therefore,

$$|A_3| = \left| \frac{B_3}{B_1} \right| \cdot |A_1|^3 \cdot |H_A G_B|^2. \tag{13}$$

It is obvious that for stable operation of the predistorted transmitter, the magnitude of the transfer coefficients, $|A_1|$ of the predistorter and $|H_A G_B|$ of the interconnecting networks (which includes the up converter), must be very well controlled and remain stable over time and temperature. The same is true for the gain of the TWT, represented by $|B_1|$ in eq. (13).

The phase condition for the upper sideband is

$$(\sphericalangle A_3 - \sphericalangle A_1) = \pi + (\sphericalangle B_3 - \sphericalangle B_1) \qquad \text{(USB)}, \tag{14a}$$

and for the lower sideband is

$$(\sphericalangle A_3 - \sphericalangle A_1) = \pi - (\sphericalangle B_3 - \sphericalangle B_1) \qquad \text{(LSB)}. \tag{14b}$$

Thus, means have to be provided in the predistorter to achieve both phases since either the USB or LSB may be deployed, depending on the channel, in the AR6A System.

2.4 Interaction distortions

In the derivation of the overall nonlinear transfer function for the transmitter, terms involving convolutions of higher than third order have been neglected. It is worthwhile, however, to discuss the next higher-order term, which is found to be

$$D_5 = 3B_3(H_A G_B A_1 X)^{2*} * (H_A G_B A_3)(X)^{3*}. \tag{15}$$

Assuming that the transmitter is aligned properly, such that conditions in eqs. (11) and (12) are fulfilled, the magnitude of D_5 becomes

$$|D_5| = 3|H_A G_B A_1|^5 \cdot \left| \frac{B_3^2}{B_1} \right| \cdot |X^{5*}|. \tag{16}$$

We see from the above expressions that this fifth-order distortion is strictly due to interaction between the third-order terms of the predistorter and TWT and that the magnitude is very sensitive to the first- and third-order coefficients of the TWT. For $|D_5|$ to remain small, $|B_1|$ (i.e., TWT gain) should be large and $|B_3|$ of the device must be small at the outset.

2.5 Residual third-order distortions

In practice it is not possible to fulfill the conditions given by eqs. (11) and (12) at all frequencies. It becomes necessary, therefore, to estimate the amount of residual distortions after imperfect predistortion. In particular, the effect of deviations from the ideal condition in eq. (11) is important because of the strong sensitivity of D_3, defined in eq. (10), to the transfer function $T = H_A G_B A_1$. We will, therefore, attempt to analyze $|D_3|$ as a function of small deviations δ_c and δ_α of $|T|$ and $\alpha = \sphericalangle T$, respectively. Thus, we express T as

$$T = (C + \delta_c) \cdot e^{-j(\omega\tau + \delta_\alpha)}, \tag{17a}$$

which is approximately equal to

$$T \simeq Ce^{-j\omega\tau} + (\delta_c/C - j\delta_\alpha)Ce^{-j\omega\tau}$$

or

$$T \simeq T_o + E \cdot T_o. \tag{17b}$$

Obviously, $T_o = C \cdot e^{-j\omega\tau}$ represents the ideal transfer function, which we will assume to fulfill the condition in eq. (11). The error term $E \cdot T_o$ is proportional to the complex quantity

$$E = \frac{\delta_c}{C} - j\delta_\alpha, \tag{18}$$

and is a function of frequency. Introducing eq. (17b) into eq. (10) we obtain

$$D_3 = B_1 \frac{A_3}{A_1} (T_o + ET_o)(X)^{3*} + B_3(T_oX + ET_oX)^{3*}. \tag{19}$$

Performing the convolutions on the last term, eq. (19) can be expressed as

$$D_3 = B_1 \frac{A_3}{A_1} T_o(X)^{3*} + B_3C^2T_o(X)^{3*} + D_{3R}. \tag{20}$$

With the predistorter active and conditions in eqs. (11) and (12) being met, the first two terms cancel. The residual distortions are then approximately given by

$$D_{3R} = B_1 \frac{A_3}{A_1} ET_oX^{3^*} + 3B_3(T_oX)^{2^*} * (ET_oX) + \cdots, \tag{21}$$

where terms of higher order in (ET_oX) have been neglected. By introducing eq. (12) and using the identity $(T_oX)^{2^*} = CT_o(X)^{2^*}$ we obtain

$$D_{3R} = -B_3C^2T_o[EX^{3^*} - 3X^{2^*} * (Ee^{-j\omega\tau}X)]. \tag{22}$$

The power spectrum of D_{3R} for given deviations E can be computed using eq. (22). Unfortunately, this involves, in general, complex numerical calculations and results do not easily provide further insight. Instead, we have used eq. (22) to arrive at an upper bound for the magnitude of D_{3R}. By using the inequality

$$|D_{3R}| \leq |B_3| \cdot C^3[|E| \cdot |X^{3^*}| + 3|X^{2^*}| * |E \cdot X|] \tag{23}$$

we have examined each term within the brackets individually for two typical, yet simplified, error functions E.

2.5.1 Case 1—Out-of-band residual distortions

If we assume that $|E|$ and $|X|$ are as illustrated in Fig. 4, it becomes obvious that only the first term in eq. (23) contributes to residual distortions, because the product $E \cdot X$ vanishes everywhere. Typically, the bandpass filter between converter and TWT produces this type of residual out-of-band third-order distortion. In the AR6A System, their

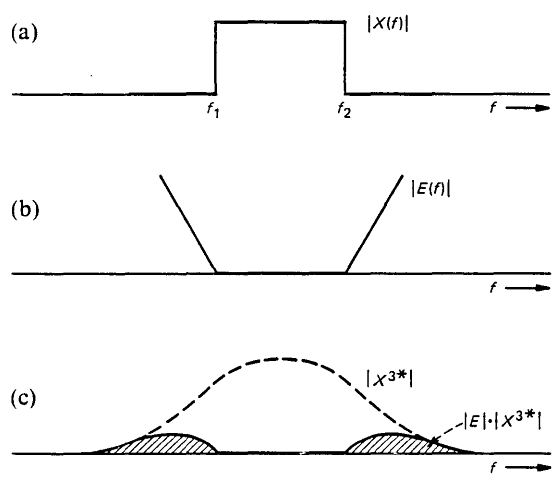

Fig. 4—Out-of-band distortions with perfect in-band predistortion: (a) input spectrum; (b) deviation from flatness; (c) distortion spectrum.

level is sufficiently small to cause no interference with other radio channels.

2.5.2 Case 2—In-band residual third-order distortions

This important case deals with parabolic in-band deviations E, as illustrated in Fig. 5. Satisfactory, but imperfect, tuning of the RF bandpass filter or insufficient flatness of the IF driver amplifier in the converter are primary causes for parabolic-type residual deviations. For simplicity, we approximate these by a triangular distribution, which causes the in-band residual distortions as shown shaded in Figs. 5c and 5d. It is important to recognize that the level of distortions due to the second term of eq. (23) varies in direct proportion to the fractional bandwidth Δf over which $|E|$ assumes a substantial magnitude. This is in contrast to the contribution stemming from the first term, the maximum level of which is only proportional to the magnitude of E. As a rule of thumb, the first term dominates if

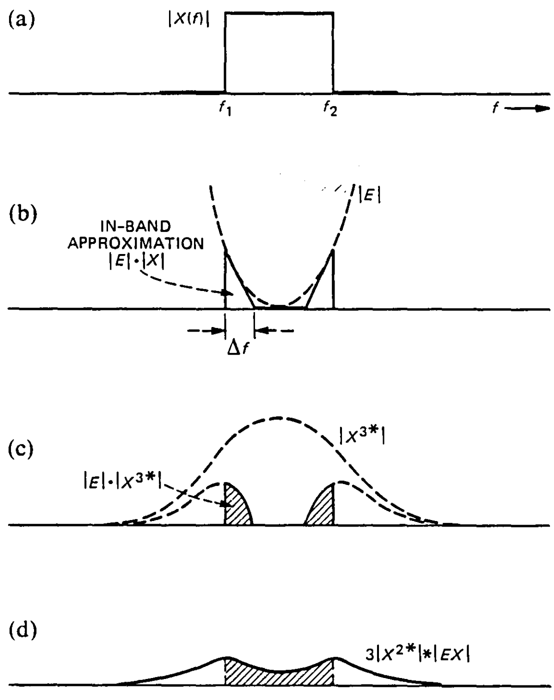

Fig. 5—Residual in-band third-order distortion spectrum: (a) input spectrum; (b) deviation from flatness; (c) residual distortion spectrum, first term, eq. (23); (d) residual distortion spectrum, second term, eq. (23).

$$3\Delta f < \tfrac{1}{4} \cdot BW,$$

where BW is the bandwidth of the input spectrum, i.e., $BW = f_2 - f_1$. In the AR6A System a special slope equalizer has been incorporated into the driver amplifier of the up converter to ensure that essentially only parabolic residual deviations remain at the very edges of the signal band, thus, keeping Δf as small as practicable.

The maximum tolerable amount of these deviations has been derived from eq. (23) by requiring that the ratio $|D_{3R}/D_{3o}|$ is below a specified level, where $|D_{3o}| = |B_3|C^3|X^{3^*}|$ represents the unpredistorted intermodulation spectrum within the message band. Thus, this ratio is a direct measure of distortion reduction. The predistortion improvement (PDI) in decibels is given by

$$PDI = -20 \log|D_{3R}/D_{3o}| \qquad \text{(dB)}. \tag{24}$$

If only the first term in eq. (23) contributes to residual distortions, the predistortion improvement becomes

$$PDI = -10 \log\left[\left(\frac{\delta_c}{C}\right)^2 + \delta_x^2\right] \qquad \text{(dB)}.$$

Figure 6 shows plots of constant PDI values as a function of phase deviation (in degrees) and flatness, defined as 20 log(1 ± δ_c/C). We see that to achieve a PDI ⩾ 40 dB, the flatness must be better than ±0.06 dB and phase linearity is required to be within ±0.4 degree, provided both deviations contribute equal amounts.

Deviations in other coefficients describing the transmitter model, in particular A_1, A_3 and B_1, B_3, have been treated similarly. Except for A_1, their effect is less complicated because they are not convolved. It can be shown that as long as the residual distortions are caused only by nonconvoluted terms, the total amount of residual distortions is bound by the inequality

$$\left|\frac{D_{3R}}{D_{3o}}\right| \leqslant \sqrt{\left(\sum_{i=1}^{N} |m_i|\right)^2 + \left(\sum_{i=1}^{N} |\Delta\psi_i|\right)^2}, \tag{25}$$

where m_i represents the deviation from flat magnitude and $\Delta\psi_i$ the deviation from phase linearity of the ith contributing coefficient. This upper bound estimate given by eq. (25) is very conservative because it is based on the assumption that at any frequency all deviations of the different component parts are causing in-phase addition. This is, in general, not true and a more useful estimate based on statistical addition is

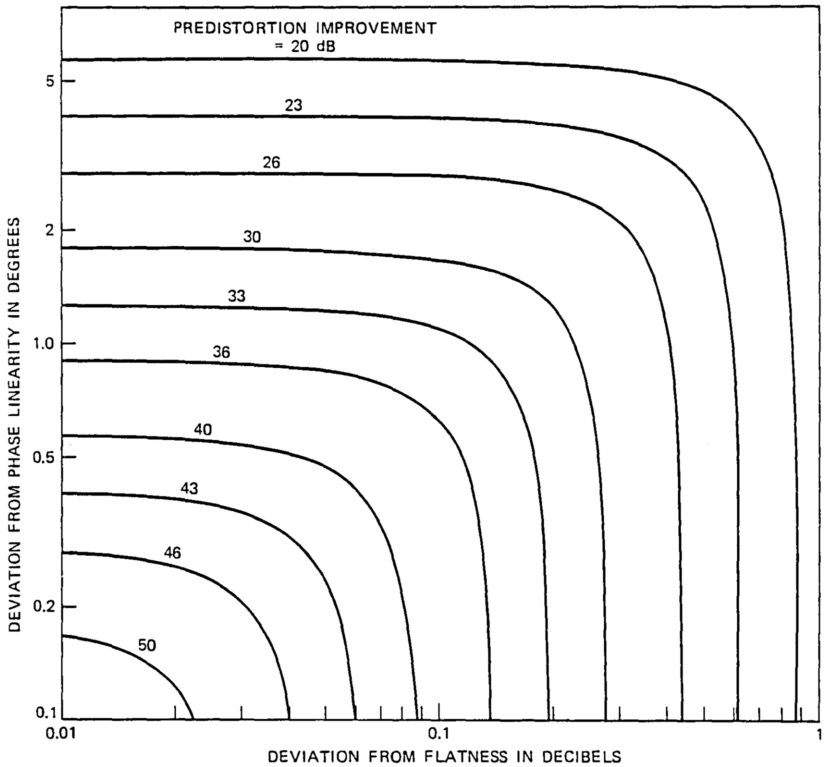

Fig. 6—Contours of constant predistortion improvement.

$$\left| \frac{D_{3R}}{D_{3o}} \right| \equiv \sqrt{\sum_{i=1}^{N} |m_i|^2 + \sum_{i=1}^{N} |\Delta\psi_i|^2}. \tag{26}$$

For the AR6A transmitter, under nominal testing conditions, the objective was to achieve a PDI of at least 30 dB over the full 30-MHz band. Allowing an additional margin of 3 dB and allocating equal deviations to each of the five possible contributors in the transmitter model (namely, A_1, A_3, B_1, B_3, and the product H_AG_B) results in a flatness requirement within ±0.06 dB and phase linearity requirement within ±0.4 degree. (Using the conservative upper bound would yield ±0.03 dB and ±0.18 degree as requirements on the model coefficients.) Since the coefficients A_1 and A_3 are only representative of the various components in the predistorter, its actual circuit requirements have been set at nominally ±0.02 dB for flatness. The corresponding requirements for phase linearity are then automatically met because of the minimum phase property of the circuits used.

III. PRACTICAL DESIGN CONSIDERATIONS

3.1 Components

The predistorter comprises components operating over the 30-MHz bandwidth centered at 74.1 MHz. The predistorter is described in the block diagram shown in Fig. 7. The key component is the nonlinear generator referred to as the cuber.[8] It consists of four back diodes arranged in a bridge configuration so that signals incident at the input of the network are attenuated by approximately 60 dB at the output. The amplitude of each third-order complementary distortion term at the cuber output is related to the amplitudes of the input signals at frequencies A, B, and C by

$$P_{A+B-C} = 46.6 + P_A + P_B + P_C \qquad \text{(dBm)}. \tag{27}$$

The driver amplifier at the input to the predistorter provides gain so that the input signal to the cuber is at the optimum level. The optimum is dictated by choosing the drive such that the cuber does not generate fifth-order distortion terms that would exceed the level assumed in eq. (9). The distortion amplifier provides gain so that the complementary distortion terms generated by the cuber have the required amplitude. The gain in the distortion amplifier must be limited because of the addition of thermal noise back into the main path. The distortion adjustment potentiometer in the output coupler is used to fine tune the amplitude to provide the exact value of $|A_3|$ required for each individual transmitter.

The phase of the third-order complementary distortion terms generated in the cuber is established by the phase resolver network.[9] This network is designed to provide a constant-phase linear-phase difference between the two output ports and has the adjustment capability to cover the range of phase angles required. The output ports of the

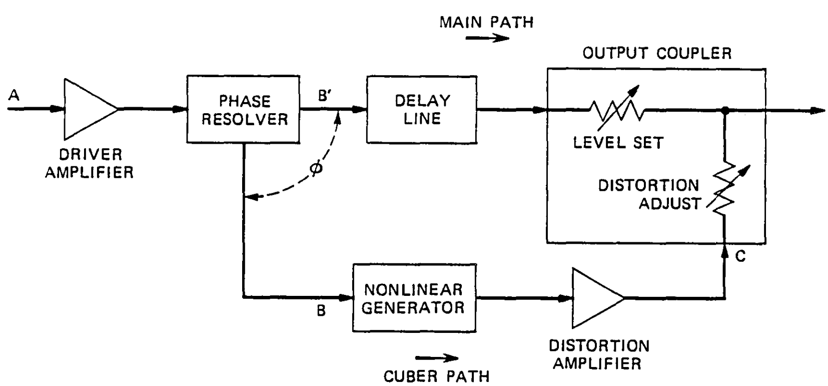

Fig. 7—Predistorter block diagram.

resolver network are arranged so that the connections to these ports can be interchanged. By doing so, the predistorter is adaptable for either USB or LSB operation.

The delay line in the main path is added to match the delay of the cuber path. To account for tolerances of the components, a delay adjustment is required during the manufacturing procedure so that the delays of the two paths are matched as closely as possible.

Because of the large amount of delay required, mismatches at the ends of the delay line result in shape over the 30-MHz channel that is difficult to equalize. Thus, the return loss of the appropriate ports of the delay line, phase resolver, and output coupler is carefully controlled.

3.2 Equalization

Transmission deviations from a flat shape in either the main or cuber path result in residual third-order distortion. Two slope equalizers in the predistorter are employed to control the linear portion of the deviations.

The first slope equalizer adjusts the path from point A to point B, as shown in Fig. 7. This slope equalizer is part of the driver amplifier. The end points over the 30-MHz channel are equalized within ± 0.02 dB. This procedure imposes strict shape requirements on the resolver network so that

$$|H_{AB}| \cong |H_{AB}'|. \tag{28}$$

The second slope equalizer adjusts the path from point B to point C. This slope equalizer is part of the distortion amplifier. The equalization process is accomplished by first disabling the bridge network in the cuber, then equalizing the end points over the 30-MHz channel to within ± 0.02 dB.

3.3 Alignment of transmitter

Since each predistorter is equalized during the manufacturing process as previously discussed, the rest of the transmitter must be equalized after all the parts are assembled. From eq. (11) these separate equalizations are equivalent to requiring that H_A be aligned in the transmitter configuration while $G_B A_1$ is aligned separately.

From Figs. 2 and 3, we see that H_A includes the amplifier/modulator, bandpass filter, and the input network of the TWT. A slope equalizer is incorporated in the amplifier/modulator. It is clear that the interface of this portion of the transmitter is somewhere within the active portion of the TWT and is, therefore, not accessible using standard methods. A special scanning intermodulation (IM) test set[10] was designed to permit this measurement to be made. The scanning capability of the test set permits the transmission path to be slope equalized over the 30-MHz channel.

The overall result of the equalization process is measurable using the scanning IM test set to measure the predistortion improvement across the band. The results of a number of such measurements are summarized in Fig. 8. Two types of TWTs have been used in the AR6A transmitter. For the transmitters using the WE 473A TWT, the average PDI for 24 TWTs indicates a minimum value of 33 dB. To obtain this result, the deviation from flatness must be approximately 0.1 dB. For the transmitters using the KS-22469 TWT, the average PDI for 27 TWTs indicates a minimum value of 35 dB with the corresponding deviation from flatness being approximately 0.07 dB.

Final measurements are made using the scanning IM test set on each transmitter. At this time approximately 3000 units have been produced. The results still show an average minimum PDI of approximately 35 dB.

IV. PERFORMANCE

4.1 Variation with time

Predistortion of the traveling-wave tube has been demonstrated in the laboratory and manufacturing environment. Transmitter perform-

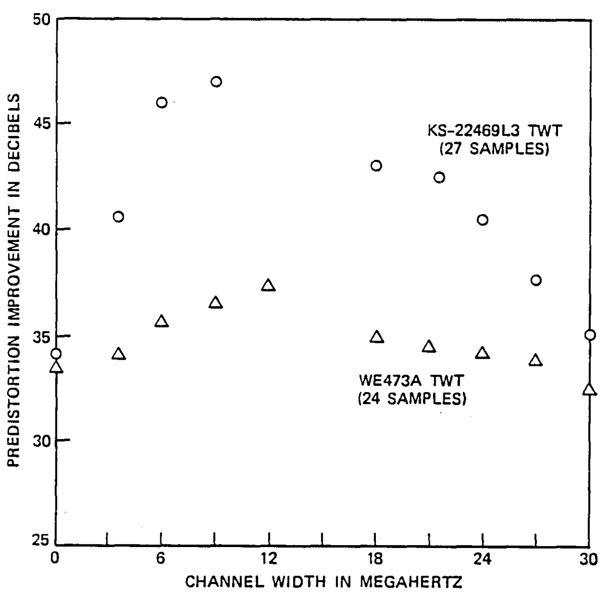

Fig. 8—Average predistortion improvement across a channel.

ance in the field with time and temperature is being tracked by observing the AR6A System in its operating environment.

Data accumulated over a period of several years for 25 TWT amplifiers indicate that the performance with time largely depends upon the stability of the TWT and power supply. The change of gain of the TWT amplifier has been, in general, less than 0.35 dB per year (corresponding to a PDI of 22 dB). Exceptions exist; some amplifiers have demonstrated very large (>2 dB) gain changes. Although no regular monitoring plan has been established, occasional evaluation of selected channels has corroborated the stability estimates.

4.2 Variation with temperature

Temperature changes also cause transmitter gain variation and, therefore, cause the PDI to vary. The AR6A transmitter was tested over an ambient temperature range from 40 to 120°F. These tests revealed that the variation of the PDI was due to transmitter gain changes. The largest contributor to the gain change is the TWT amplifier. Separate tests of five amplifiers established an average change of 0.35 dB over the temperature range. For a transmitter aligned at 80°F, this type of gain variation results in a worst-case PDI of 27 dB.

4.3 Summary

In some cases, gain variations with time and temperature significantly exceed the typical value expected. The AR6A System design compensates for these extreme cases by allowing sufficient margin. The overall intermodulation level in a switch section is monitored. When limits are exceeded, the monitoring system can identify the transmitter or transmitters that require adjustment. In a typical switch section, the random aging and temperature conditions result in an overall switch section performance that satisfies objectives.

V. ACKNOWLEDGMENTS

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APPENDIX

On the Transfer Function for the Up Converter With a Traveling-Wave Tube

The conversion characteristic of a Schottky diode converter is, for our purposes, most conveniently approximated by the ideal multiplier function

$$z(t) = \mathrm{K}\cdot s(t)\cdot z_{\mathrm{m}}(t), \tag{29}$$

where K is a frequency-independent constant, $s(t)$ the local oscillator (pump) signal, and $z_{\mathrm{m}}(t)$ the modulating (IF) signal. With $s(t) = \cos \omega_o t$, one can see that eq. (29) represents a double-sideband signal with suppressed carrier. Since the bandpass filter following the converter in the actual transmitter suppresses the unwanted sideband, as well as any residual carrier, the use of eq. (29) is indeed justified. Applying a single-tone modulating signal $z_{\mathrm{m}}(t) = A_{\mathrm{m}}\cos(\omega_{\mathrm{m}}t + \phi_{\mathrm{m}})$, one obtains from eq. (29) the lower single-sideband (SSB) signal

$$z(t) = \frac{A_{\mathrm{m}}}{2}\,\mathrm{K}\;Re\{e^{j(\omega_o-\omega_{\mathrm{m}})t}\cdot e^{-j\phi_{\mathrm{m}}}\}, \tag{30a}$$

and the upper SSB signal

$$z(t) = \frac{A_{\mathrm{m}}}{2}\,\mathrm{K}\;Re\{e^{j(\omega_o+\omega_{\mathrm{m}})t}\cdot e^{+j\phi_{\mathrm{m}}}\}. \tag{30b}$$

The converter, thus, not only translates the modulating frequency to the desired RF band, but also causes the conjugation of the modulating signal phase. One may, therefore, define two ideal converter transfer operations linking the two frequency domains as follows:

$$Z_{\mathrm{LSB}} = Z(\omega_o - \omega_{\mathrm{m}}) = \mathrm{K_c}\cdot\check{Z}_{\mathrm{m}}(\omega_{\mathrm{m}}) \qquad \text{(LSB)} \tag{31a}$$

and

$$Z_{\mathrm{USB}} = Z(\omega_o + \omega_{\mathrm{m}}) = \mathrm{K_c}\cdot Z_{\mathrm{m}}(\omega_{\mathrm{m}}) \qquad \text{(USB)}, \tag{31b}$$

where $\check{Z}$ is the complex conjugate of Z, and $\mathrm{K_c}$ a frequency-independent constant. Note that these operations describe the combined functions of up converter and bandpass filter. Introducing eq. (31a) into the nonlinear transfer function, eq. (4) of the main text, yields (with some simplifications)

$$Y_{\mathrm{LSB}} = H_B(B_1 H_A \check{Z}_{\mathrm{m}} + B_3(H_A \check{Z}_{\mathrm{m}})^{3^*}) \tag{32a}$$

and

$$Y_{\mathrm{USB}} = H_B(B_1 H_A Z_{\mathrm{m}} + B_3(H_A Z_{\mathrm{m}})^{3^*}). \tag{32b}$$

With these expressions we have derived nonlinear transfer functions for the cascade of up converter, bandpass filter, and TWT. Note that for the LSB and the USB case, the modulating signal is Z_{m} and it is the converter that performs the conjugation. It also must be mentioned that for the intentions of this model, the converter constant $\mathrm{K_c}$ and the transfer function of the bandpass filter have tacitly been absorbed in H_A.

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The AR6A Single-Sideband Microwave Radio System:

System Networks

By R. L. ADAMS,* J. L. DONOGHUE,* A. N. GEORGIADES,* R. E. SHEEHEY,* J. R. SUNDQUIST,* and C. F. WALKER*

(Manuscript received February 14, 1983)

The paper describes some of the key networks needed for the AR6A System for the predistorter, Intermediate Frequency (IF) filtering, and terminal multiplex. The design of a passive broadband adjustable phase splitter for the predistorter is described. Network designs using inductors, capacitors, and, in some cases, crystals for IF filtering, equalization, and multimastergroup connection are also described.

I. INTRODUCTION

A large number of filters and equalizers, and an adjustable phase splitter are needed in AR6A for pilot insertion and removal, band limiting, amplitude correction, and predistortion circuitry in repeaters and in multiplex equipment. Technology developed for TD and TH radio[1,2] and L4 and L5 coaxial-carrier transmission systems[3–5] was enhanced to meet the specific and demanding needs of the AR6A† System filters and equalizers; however, a special approach was required for the design of an adjustable phase resolver (phase splitter) for the predistorter. The phase resolver is an adjustable Inductor Capacitor (LC)‡ network that allows matching the phase of the predistortion

* Bell Laboratories.
† 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 the *Journal.*

signal to the opposite of the phase of the distortion component of the main signal. The filters provide wider bandwidths and higher discrimination than their FM radio counterparts, and amplitude equalization, rather than delay equalization, is required. Extensive use was made of computer aids for design and analysis. A high-speed, precision interactive mini Computer-Operated Transmission Measurement Set (COTMS) proved necessary for laboratory development of networks. More than 700 models of about 140 network designs were needed for development and final design of the AR6A System.

The theory and design of the phase resolver for the AR6A predistorter is described in some detail. Descriptions of the Intermediate Frequency (IF) filter and equalizer, the upper and lower IF bandpass filters, and the multimastergroup connector filters follow.

II. PHASE RESOLVER

2.1 Function

The phase resolver is a passive three-port network used in the AR6A predistorter.[6] It splits the predistorter input signal into two equal-amplitude signals that have a constant phase difference in the IF band between 59 and 89 MHz. The phase difference is manually adjustable. This provides adjustment of the phase predistortion component to cancel the distortion products generated in the traveling-wave tube.

2.2 Theory

The phase-resolver network consists of a quadrature hybrid network, two attenuators, and a combining hybrid transformer connected as shown in Fig. 1. The outputs can be shown to be

$$\begin{aligned} E_B &= 0.5\ E_A(G_1 + jG_2), \\ E_C &= 0.5\ E_A(-G_1 + jG_2). \end{aligned} \tag{1}$$

Thus, the magnitude of either output is

$$|E_{B,C}| = 0.5\ |E_A|\ \sqrt{G_1^2 + G_2^2}, \tag{2}$$

and the phase difference is

$$\Theta_B - \Theta_C = 2\ \tan^{-1}(G_2/G_1). \tag{3}$$

A key element of this network is the broadband quadrature hybrid network.

To achieve a wideband quadrature hybrid network, a Cauer configuration[7] was chosen. This is a four-port network consisting of two hybrid transformers connected to two lossless LC networks, N and N', as shown in Fig. 2. With port A driven, ports B and C will have quadrature phase, and port D will be isolated if N and N' are

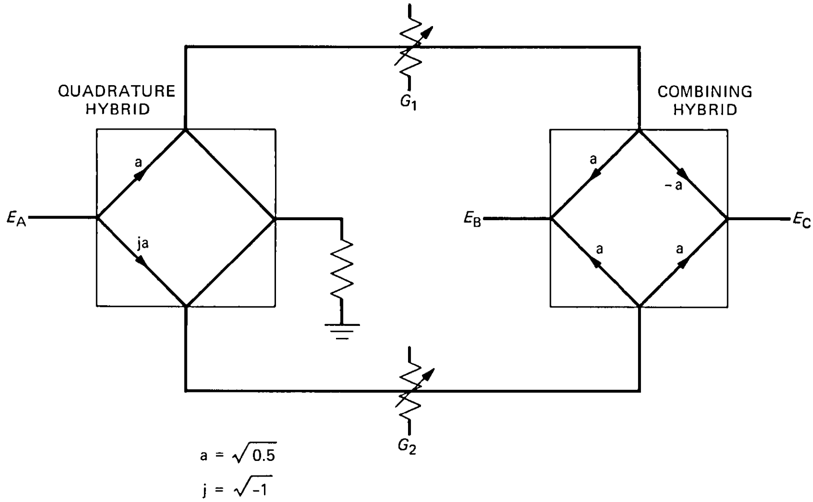

Fig. 1—Phase-resolver block diagram.

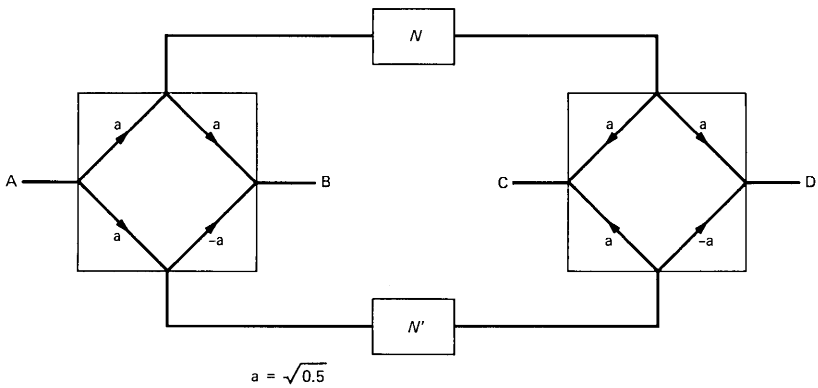

Fig. 2—Cauer network.

dual symmetric networks. This can be determined if the scattering parameters for the complete network are derived in terms of the scattering parameters of N and N'.

$$\begin{aligned} s_{AA} &= (s_{11} + s'_{11})/2 \\ s_{BA} &= (s_{11} - s'_{11})/2 \\ s_{CA} &= (s_{21} + s'_{21})/2 \\ s_{DA} &= (s_{21} - s'_{21})/2. \end{aligned} \tag{4}$$

If N and N' are dual networks, then

$$s_{11} = -s'_{11}, \tag{5}$$

and

$$s_{21} = s'_{21}.$$

Then,

$$s_{AA} = 0,$$

$$s_{BA} = s_{11},$$

$$s_{CA} = s_{21},$$

$$s_{DA} = 0. \tag{6}$$

The other twelve scattering parameters for the four-port network have a corresponding form. These relations describe a hybrid network with port A isolated from port D, and port B isolated from port C. It is a quadrature network if N and N' are not only dual but also symmetric lossless LC networks. This can be shown as follows.

It is convenient to define a split ratio, Γ, as

$$\Gamma = \frac{s_{BA}}{s_{CA}} = \frac{s_{11}}{s_{21}}. \tag{7}$$

For a quadrature network, Γ must have the following form:

$$\Gamma(s) = s\,\frac{P_{11}(s^2)}{P_{21}(s^2)} \quad \text{or} \quad \frac{P_{11}(s^2)}{sP_{21}(s^2)}. \tag{8}$$

Here, P_{11} and P_{21} are polynomials. With $s = j\omega$, phase quadrature can be observed. To avoid right-half plane transmission zeros requiring right-half plane zeros, Γ is restricted to the following form:

$$\Gamma(s) = \frac{P_n(s^2)}{s^{2m+1}}. \tag{9}$$

Here, P_n is an nth-order polynomial and m is an integer. Then the networks, N and N', are simple ladder networks.

The design of a quadrature Cauer network consists of first finding a suitable polynomial, $P_n(-\omega^2)$, such that $|\Gamma|$ approximates unity over the frequency band of interest. A corresponding symmetric ladder network and its dual are then synthesized to realize N and N'. The synthesis procedure can be simplified by taking advantage of the fact that the networks are symmetric.

2.3 Realization

Extensive use was made of computer aids and computer-operated transmission measuring sets during the design and development. The hybrid transformers were characterized on COTMS.[8] With this characterization used in a circuit analysis program, an optimization routine could be used to modify N and N' to compensate for parasitics. For model characterization and laboratory correction of parasitics, a high-accuracy, high-speed computer-operated transmission measurement set proved necessary. This set, "Mini COTMS," has a measurement speed fast enough for computer-corrected simultaneous real-time display of loss, phase or delay, and input and output return loss.

Figure 3 is a photograph of the phase resolver.

III. IF FILTER/EQUALIZER

3.1 Function

This circuit provides the 74.13-MHz center frequency IF filtering and amplitude equalization. Figure 4 shows that four networks are used. The bandpass filter limits adjacent channel interference and image interference in the Multimastergroup Translator for Radio (MMGT-R). One basic equalizer and two mop-up equalizers provide static amplitude equalization. The basic equalizer corrects for systematic amplitude variations in the overall Transmitter-Receiver (TR) bay transmission characteristic. Two mop-up equalizers correct for unit-to-unit amplitude deviations.

3.2 Design

Design was facilitated by two interactive computer programs. One does the rational function approximation and insertion loss synthesis; the other performs Norton transformations to optimize filter topology and element values. A general-purpose linear network analysis program allowed simulation of parasitic effects in any of the designs.

An 11th-order filter was synthesized to meet the objectives. Two

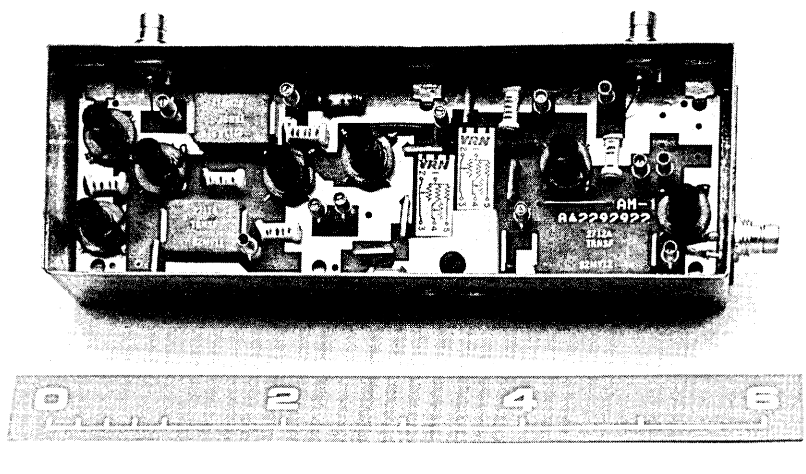

Fig. 3—Photograph of phase resolver.

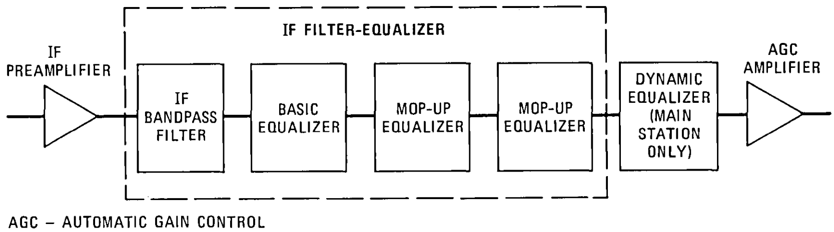

Fig. 4—IF filter-equalizer block diagram.

bridged-T sections are needed to compensate for amplitude variations caused by the finite Q of the filter inductors. The filter is assembled on a printed wiring board that is attached to a cast-aluminum housing as shown in Fig. 5. Electrostatic shields are contained within the housing. Field tests have confirmed that the filter meets system requirements.

Four mop-up equalizer designs achieve positive-slope, negative-slope, positive-quadratic-shape, and negative-quadratic-shape amplitude response. Bridged-T sections are used to realize each of the required shapes. Frequency response measurements of the AR6A switch section involved are used to choose the shapes used. When a mop-up equalizer is not needed, a 3.5-dB pad is used.

IV. LOWER AND UPPER IF BANDPASS FILTERS

4.1 Function

The transmitting-output and the receiving-input spectrum of the MMGT-R each cover the 59.844- to 88.460-MHz band and contain ten mastergroups. This frequency spectrum is split into two bands of five mastergroups each. The Lower-band Intermediate Frequency (LIF) is from 59.844 to 73.116 MHz, and the Upper-band Intermediate Frequency (UIF) is from 75.188 to 88.460 MHz.

Two bandpass filters (BPFs), the LIF and the UIF filters, are used in the MMGT-R to select each of these bands in the modulator and in the demodulator. The filters reject the carrier frequency, the upper sideband, and second and third harmonics of the carrier. Lower sideband loss keeps crosstalk and tone interference into adjacent radio channels, due to undesired modulator products, at acceptable levels.

4.2 Design

The designs of the two filters are similar, so only the LIF BPF will be covered. The stopbands cover a wide frequency range and very high losses are required. To achieve reasonable element values, to reduce the effect of parasitics for this wideband, and to ease tuning, a low-pass filter and a high-pass filter in tandem are used rather than a

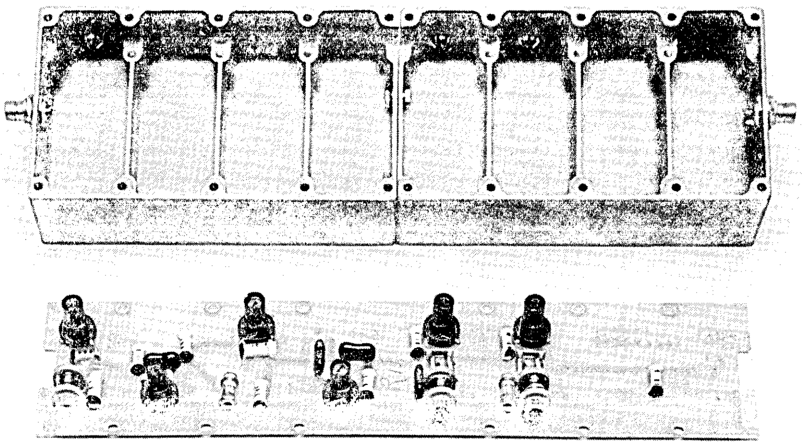

Fig. 5—Photograph of IF filter.

bandpass design. Computer-aided insertion loss methods (Section 3.2) were used to synthesize a 13th-order Cauer elliptic high-pass filter and a ninth-order equal ripple passband with arbitrary stopband low-pass filter. Norton and pi-delta transformations were applied to achieve equal termination resistances and desirable element values. Two bridged-T constant resistance sections are required to compensate for amplitude variations in the passband caused by the finite Q of the inductors.

The filter is assembled on two printed wiring boards and housed in a drawn steel enclosure (see Fig. 6). One board contains the equalizer, pad, and high-pass sections; the other board contains the low-pass section. Each filter inductor is shielded, and there is a shield between the printed wiring boards to reduce magnetic coupling between inductors.

V. MULTIMASTERGROUP CONNECTOR FILTERS

5.1 Function

The AR6A multimastergroup (MMG) connector is a precision passive network which allows a direct connection from the MMGT-R receiver to the MMGT-R transmitter when separation of the various components of the MMG signal is not required for routing or other purposes. In this application, the type-B Mastergroup Translators (MGTB) and other multiplexing equipment are not required. The key elements of the MMG connector are a highly selective bandpass filter and a crystal band-elimination filter. Figure 7 is a photograph of the two filters. The bandpass filter selects the 8.628- to 21.9-MHz MMG

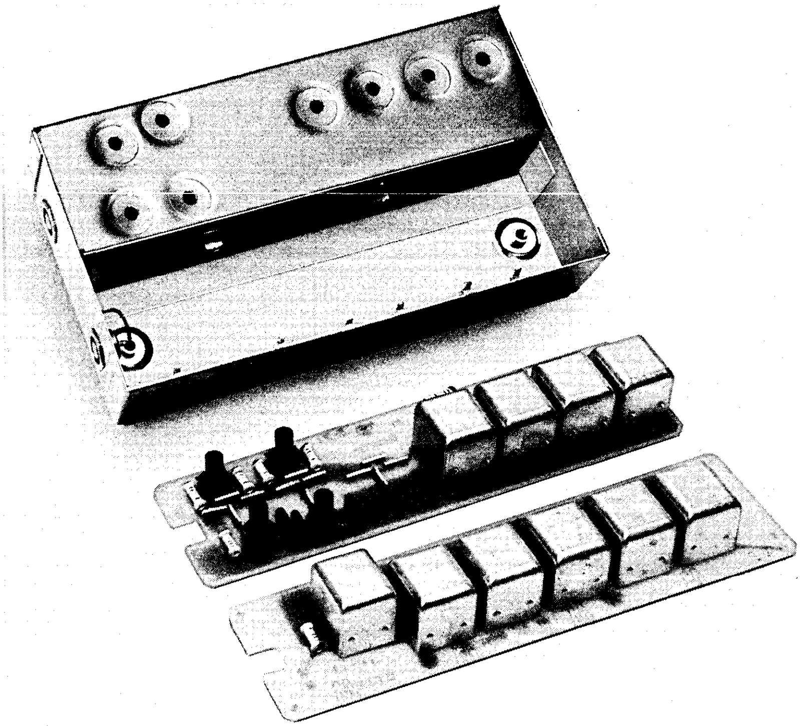

Fig. 6—Photograph of LIF BPF.

spectrum and rejects other frequencies to avoid crosstalk between radio channels. It also suppresses two pilot tones. The Band-Elimination Filter (BEF) suppresses four additional pilot tones.

The filters must have minimum effect on the message spectrum in spite of the high selectivity required. Wide skirts on the pilot suppression characteristic are needed to reduce noise near the pilots. The skirts increase the complexity of the crystal BEF significantly.

5.2 Design

The electrical design of the bandpass filter was achieved with the help of two interactive computer programs (Section 3.2). The inductor values were chosen to match a family of precision high-Q adjustable inductors originally designed for MGTB.[5] A complex arm bridged-T equalizer section was added to compensate for loss variations due to finite Q effects. The result is a 22nd-order loss-equalized bandpass filter. It was also possible to suppress two of the pilots by placing six crystal units at each frequency in shunt arms of the filter.

Four filters in tandem are used to suppress the remaining four pilots. For each frequency, a low-pass filter is designed to pass the desired

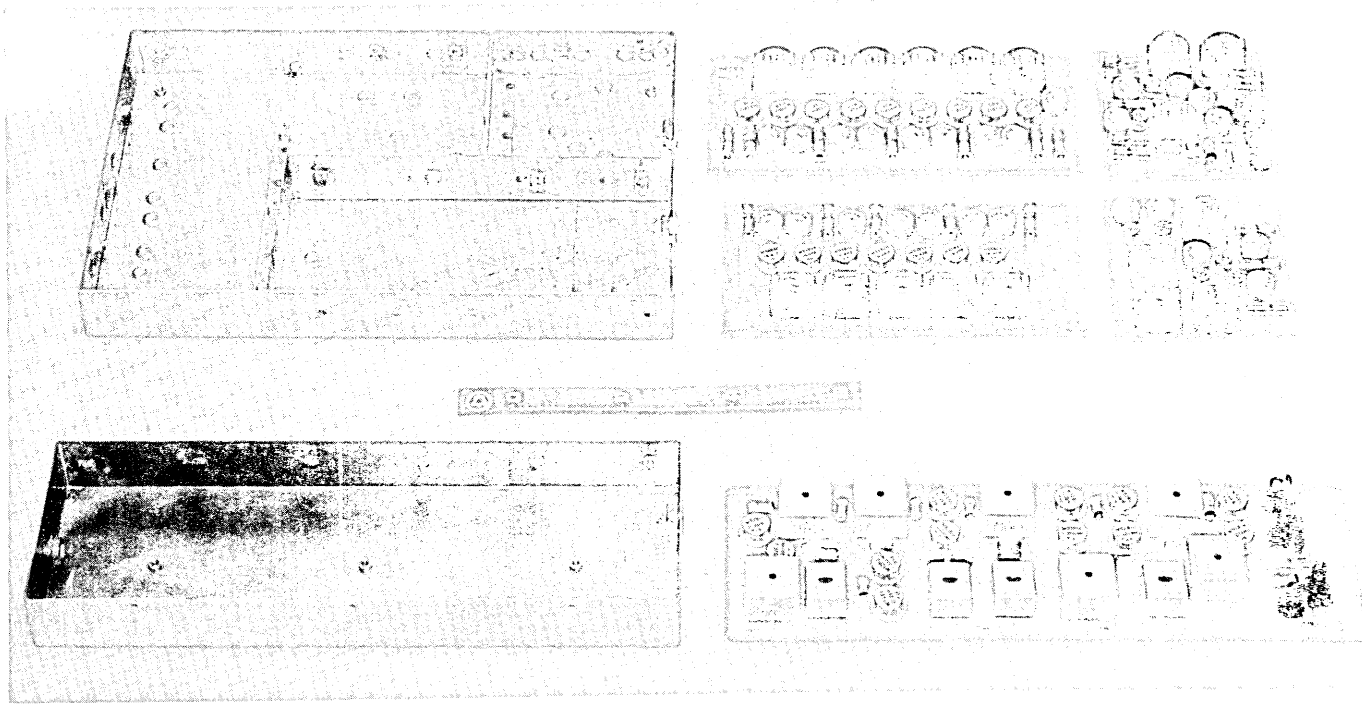

Fig. 7—Photograph of MMG connector filters with bandpass filter at top and crysal band-elimination filter at bottom.

spectrum, and crystal units are placed in shunt arms to reject the pilot. Because of the number of crystal units required to achieve the desired suppression characteristic, separate low-pass filters are needed for each frequency. A total of 28 crystal units are used in the band-elimination filter in addition to the 12 crystal units in the bandpass filter.

VI. SUMMARY

The paper has described the design of some of the key filters, equalizers, and networks for AR6A. The authors would like to thank S. Darlington for his innovative Cauer coupler designs, and F. J. Witt, T. H. Simmonds, R. P. Snicer, and J. L. Garrison for their inspiration and guidance.

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The AR6A Single-Sideband Microwave Radio System:

Special Test Equipment

By R. I. FELSBERG* and M. E. SANDS*

(Manuscript received January 22, 1983)

This paper describes two specialized test sets designed specifically for AR6A transmit-receive radio equipment alignment. The scanning intermodulation (IM) test set can adjust the transmitter for proper gain shape and optimum predistorter performance. It can also confirm that the transmitter is meeting its IM requirements. The pilot selection test set is used to test the radio intermediate frequency receiver units for proper operation during changing received signal conditions. This test set provides signals to simulate either flat or selective fades and can be used to check the pilot resupply function or space-diversity switch transition levels.

I. INTRODUCTION

It was realized early in the development of the AR6A† Radio System that special test equipment would be needed to ensure that performance objectives of this new single-sideband (SSB)‡ system were met. Two specialized test sets have been designed for this purpose: (1) the scanning intermodulation (IM) test set, and (2) the pilot selection test set. Both sets are self-contained in carrying cases with handles and

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* Bell Laboratories.

† Amplitude Modulation Radio at 6 GHz for the initial (A) version.
‡ Acronyms and abbreviations used in the text and figures of this paper are defined at the back of the *Journal.*

are provided with the necessary cables and other accessories for use in the field.

The scanning IM test set has been developed to adjust the transmitters in the AR6A Radio System to meet the critical linearity requirements necessary for a multihop radio route. Swept wideband measurement of IM single-sideband amplitude-modulated (SSBAM) radio is required for predistorter and up-converter alignment and to verify that the high degree of linearity necessary for this type of transmission has been obtained. The transmitter alignment is accomplished by one single-tone level measurement/adjustment and three separate three-tone, third-order IM measurement/adjustments made with the test set. An auxiliary oscilloscope (the display section of the radio transmission test set) is used to display the resulting transmitter intermodulation characteristics over the frequency band of interest.

The AR6A receiver conditions the received signal prior to retransmission to the next station. The circuits that perform this function require special test signals to verify the correct functioning of this portion of the AR6A System. The pilot selection test set was developed to supply the required signals in a convenient manner.

II. SCANNING INTERMODULATION TEST SET

2.1 General

The single-sideband amplitude-modulated transmitters in AR6A are required to be extremely linear so that the third-order IM noise that accumulates over a long route is held to an acceptable level. Transmission with very low IM noise is accomplished by using a traveling-wave tube (TWT) operated well below its maximum power output capability and by the use of predistortion. Use of the predistorter requires flat transmitter third-order IM characteristics and precise alignment to obtain maximum third-order IM reduction. Predistortion operates by generating third-order IM noise that is equal in amplitude and opposite in phase to the third-order IM noise generated by the nonlinearities in the transmitter. These nonlinearities result in cancellation (nulling) of the IM noise. To obtain wideband predistortion, the overall third-order IM from the transmitter must have a flat IM frequency characteristic. The IM test set is used to adjust the up converter and predistorter in the transmitter for the required third-order IM flatness and third-order IM distortion cancellation. In addition to the linearity requirement, the transmitter output power level is required to be held within certain limits. The alignment of the transmitter requiring IM slope adjustment and predistorter IM null adjustment interacts with the transmitter gain. As a result, the output power must be monitored as adjustments are made to maintain the required power level.

A portable scanning IM test set has been developed to provide easy and rapid alignment of transmitters in the field where the three-tone, third-order IM distortion product from the transmitter can be rapidly scanned and displayed over the entire frequency band of interest. The set is divided into two modules (Figs. 1 and 2) to reduce the weight

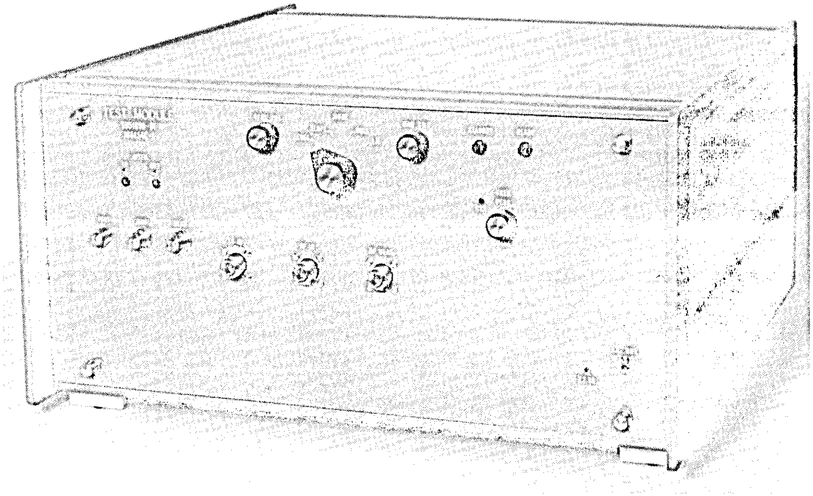


Fig. 1—Test module of scanning IM test set.

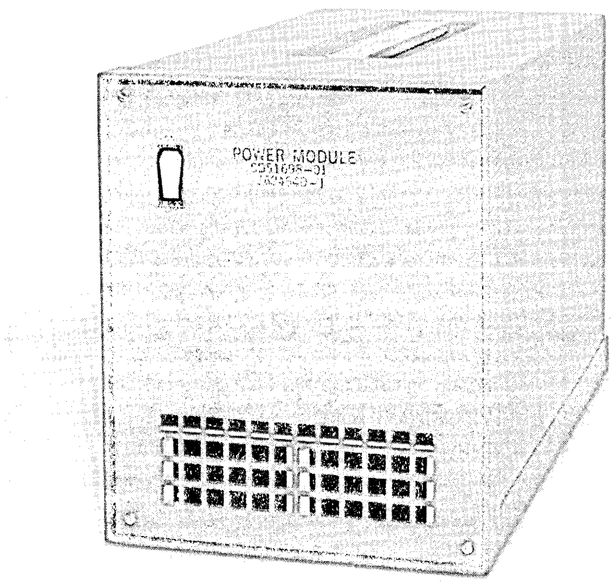


Fig. 2—Power module of scanning IM test set.

and size to acceptable values. The units are housed in metallic cases with retractable carrying handles. Separate lightweight soft-pack and hard-pack shipping cases are available to protect the modules during transportation. Space has been provided in the power module soft-pack and hard-pack to store the ten cables required for operation of the test set. An auxiliary oscilloscope is required for use with the test set to display the IM characteristics as a function of frequency. An auxiliary power meter is also required to monitor the output power during alignment.

2.2 Principle of operation

The test set can be divided into two basic sections consisting of a tone-generating section and a receiving- or signal-processing section. Referring to the simplified block diagram of Fig. 3, the three-tone generating section is located on the upper left. A manually swept, voltage-controlled oscillator (VCO) tone is split and applied to two mixers. Also applied to the two mixers are tones from crystal oscillators with frequencies that differ by 1 MHz. The two down-converted tones fall within, and can be manually swept over, the required 30-MHz intermediate frequency (IF) band. In addition, a third tone fixed in level and frequency near the center of the IF band is combined with the two manually swept tones. The three tones are sampled at the output coupler and applied to the automatic gain control (AGC) unit and then to the variolosser to maintain the level of the swept tones constant across the 30-MHz band. The three-leveled tones are applied to the device under test (DUT), i.e., the transmitter. The three-tone, third-order IM distortion product P_{a-b+c} resulting from nonlinearity in the transmitter is fixed in frequency since the difference in frequency between tone (a) and tone (b) remains fixed (1 MHz apart) although the tones are swept in frequency and the third tone (c) is also fixed in frequency. As a result, the IM product, although fixed in frequency, reflects any change in IM distortion versus frequency as a change in its amplitude when the two tones are swept over the frequency band. The tones and the resulting IM products from transmitter nonlinearities are coherently down converted from 6 GHz to IF in the signaling processing circuits shown on the right of Fig. 3. The IF signals are down converted again to 315,920 Hz so that only the single IM distortion product P_{a-b+c} from the transmitter falls within the very narrow passband (50 Hz) of the monolithic crystal filters employed in the test set. As a result, all other tones and IM products are excluded, thus preventing any spurious three-tone IM from being generated in the gain stages following the filter. In addition, a signal-to-noise advantage is gained by use of the filters. This is important, since the IM product signal level at the input to the filter is very low and therefore requires high amplification. Very stable crystal oscilla-

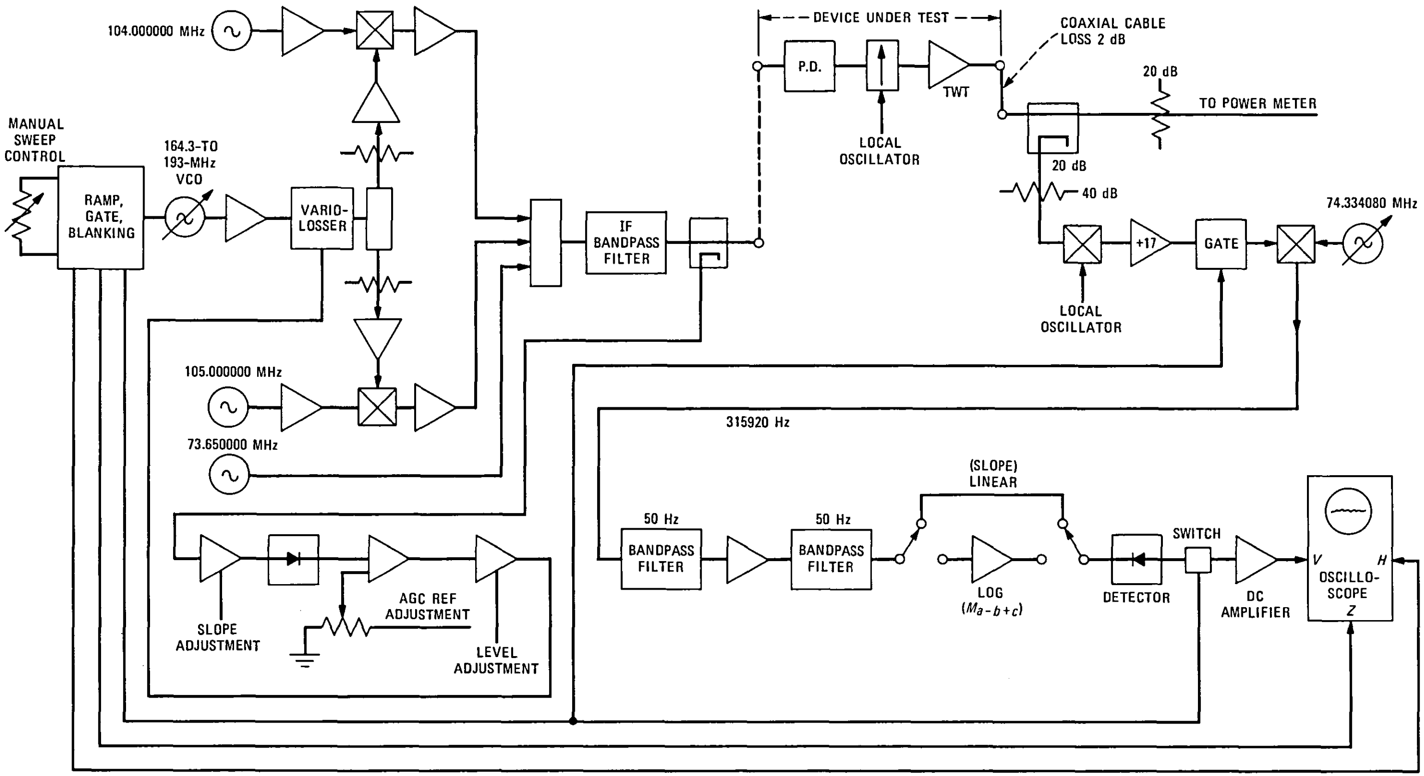
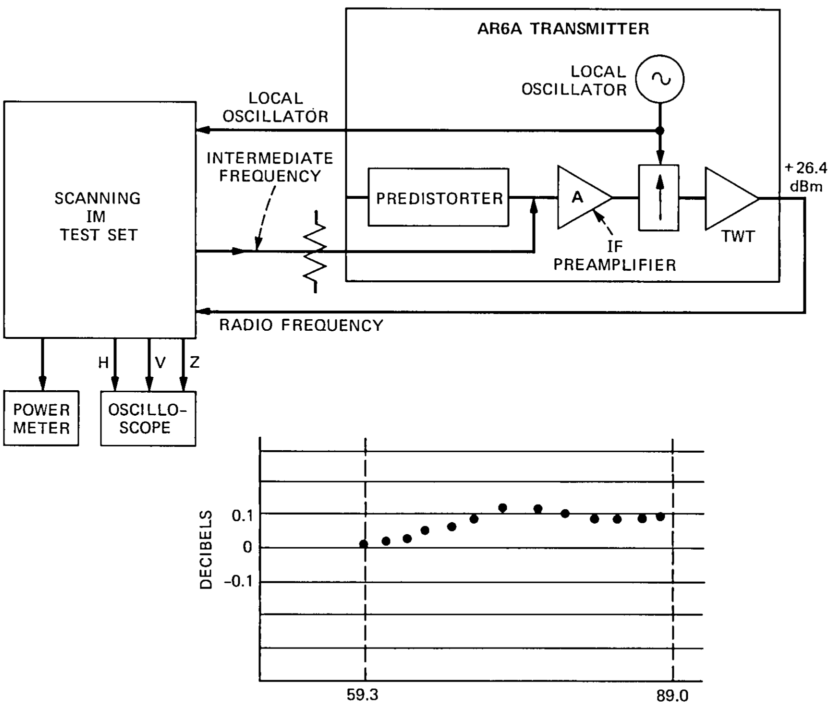


Fig. 3—Simplified block diagram of a scanning IM test set.

tors are employed to ensure that the IM product remains within the very narrow passband of the filters. Four oven-stabilized crystal oscillators are used for this purpose and have a stability of better than one part in 10^7 per day. One oscillator, employed by the signal processing section, is made adjustable by means of a knob on the front panel of the test set to center the IM product in the filter passband when necessary to compensate for crystal oscillator drift.

2.3 Tests performed

The first test (LEVEL) provides a calibrated tone to the input of the transmitter so that the power output may be set to the required level. The remaining three tests are three-tone, third-order IM distortion tests. The test set supplies three IF tones, two manually swept and one fixed in frequency, to the input of the transmitter for this purpose. In the second test (SLOPE) shown in Fig. 4, a three-tone,



SCOPE DISPLAY OF TYPICAL ADJUSTED THREE-TONE IM TRANSMITTER SLOPE OBTAINED OVER THE FREQUENCY BAND IN SLOPE TEST MODE.

Fig. 4—Transmitter slope adjustment.

third-order IM distortion product, resulting from distortion in the transmitter, is adjusted for flatness over the 30-MHz band by means of a slope adjustment control located in the up-converter preamplifier (A) in the transmitter. In the third test (NULL) shown in Fig. 5, the predistorter in the transmitter is now included in the transmitter circuit and is switched on and three tones are held fixed in frequency. The IM distortion product is minimized (nulled as shown diagrammatically in the figure as a spot moving downward) by adjustment of the phase and amplitude controls on the predistorter in the transmitter. The fourth test (IM) also shown in Fig. 5 displays typical reduction of manually swept three-tone IM distortion versus frequency obtained by predistortion. It can then be determined on the calibrated display if the IM distortion reduction meets or exceeds requirements over the entire frequency band and, if so, this indicates that the preceding tests and alignments have been satisfactorily completed.

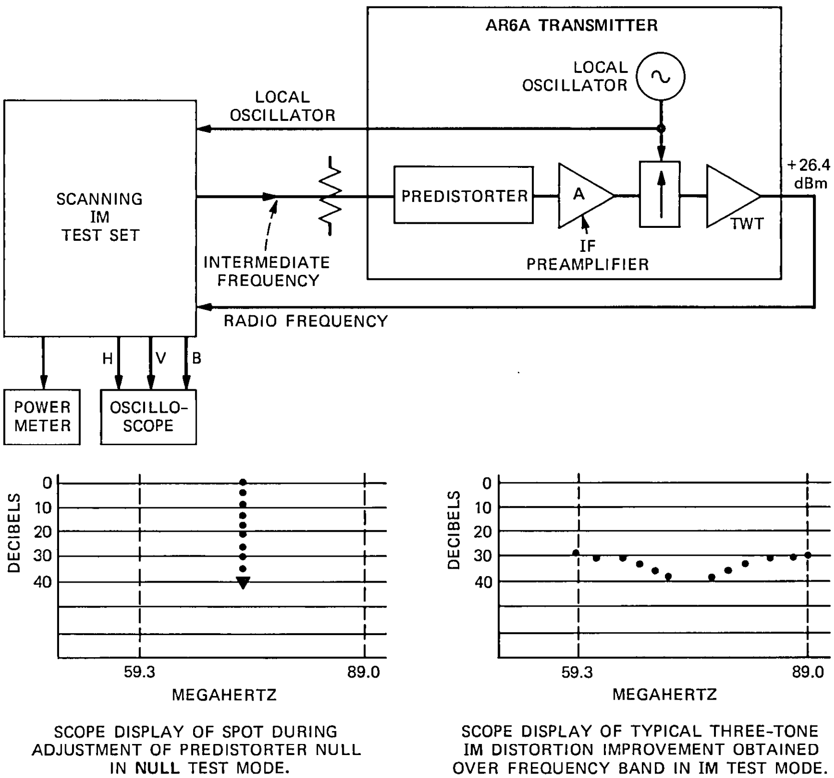


Fig. 5—Predistorter adjustment and IM improvement obtained.

2.4 Requirements for the scanning IM test set

2.4.1 Tone requirements

The test set is required to supply a single IF tone of a specified power level for the LEVEL test mode, and three tones, one fixed and two that can be manually swept in frequency, of specified power levels for the other three IM test modes. The single-tone level has been specified to have a power level equal to the full-load signal level and is used to adjust the transmitter gain to the required level. The total power of the three tones has been specified to be the same as that for the single tone. Intermodulation testing is required over the entire frequency range of each transmitter to ensure linear operation over the entire frequency band of each channel. The test set is required to produce tones that can be manually swept over the 30-MHz IF band (59.3 to 89.0 MHz). A worst-case maximum intermodulation condition has been found to exist when the three tones are held fixed in proximity to each other. The test set is designed with the three tones held fixed in frequency near the center of the band in the NULL mode so that the predistorter is nulled under worst-case conditions.

2.4.2 Spurious tone requirement

All spurious tones from the test set are required to be at least 40 dB below the power levels of the three tones. Any spurious tones that coincide with the fundamental tones as they are swept should be at least 50 dB below the power levels of the three tones so that they will add less than 0.025 dB to any of the fundamental tones. This will ensure that spurious signals will be held low enough that they will have negligible effect on the IM product. However, with swept measurements, this effect is almost unnoticeable since the spurious signals move at two or more times the rate of the fundamental tones and pass rapidly through them unnoticed, allowing the 50-dB criteria to be relaxed by 10 dB to the 40-dB requirement.

2.4.3 Tone flatness requirement

The tones from the test set are required to be flat with frequency to within 0.03 dB as they are swept over the 30-MHz frequency band. This requirement ensures that the tone levels and consequently the IM derived from these tones in the transmitter will reflect mainly the IM characteristics of the transmitter with minimum contribution of IM from the test set. The tones are set to be flat (0-dB variation) at the band edges with the maximum allowed variation of 0.03 dB at midband. The maximum variation of each of the two swept tones contributes 0.06-dB error to the CRT display of transmitter slope at midband. This small error is of little consequence, particularly since the slope can be adjusted using only the extreme frequency points.

2.4.4 *Test set IM requirements*

The three-tone, third-order IM generated by the test set is required to be at least 50 dB below that of the transmitter without predistortion. This requirement ensures that any three-tone, third-order IM originating in the test set is much less than the IM originating in the transmitter (with predistorter on) that is being measured. Therefore, the IM contributed by the test set to the measurement (less than 1 dB) would be barely discernible on a cathode ray tube (CRT) display of a 25-dB PDI (predistorter IM improvement in dB). Test set tone flatness and thermal noise from the transmitter and test set limit the amount of PDI that can be measured with reasonable accuracy to about 40 dB.

III. PILOT SELECTION TEST SET

3.1 *General*

The IF shelf of the AR6A Radio System contains a number of circuits to condition the received signal prior to retransmission. It was necessary to provide a way to check these circuits for proper operation when they were connected together in the system. A test set was designed that would condition test signals fed to the AR6A IF shelf to be representative of the type that would be experienced during normal system operation.

3.2 *Principle of operation*

The required test signal is composed of three pilot tones at frequencies of 62.448, 75.122, and 85.856 MHz. These three tones are varied in amplitude in certain prescribed ways to exercise the IF shelf signal correction circuits. This signal simulation technique checks for the proper functioning of the shelf, resupply switch, pilot detector, receiver control, AGC amplifier, and the dynamic equalizer.

3.3 *Description of test set*

The pilot selection test set (PSTS) is a portable unit that can be easily transported for field testing of the AR6A System. The unit is completely passive and requires no source of power. The PSTS must be used in conjunction with a suitable power meter.

The PSTS is made up of narrowband filters, shape networks, continuously variable attenuators, and a precision step attenuator. These components are interconnected in various ways by means of patch connectors on the front panel of the test set. All attenuators, shape networks, and filters can be used as individual items by access from the front panel jacks (see Fig. 6).

The PSTS does not contain the oscillators that supply the three required pilot tones (see Fig. 7). Each AR6A support bay contains a

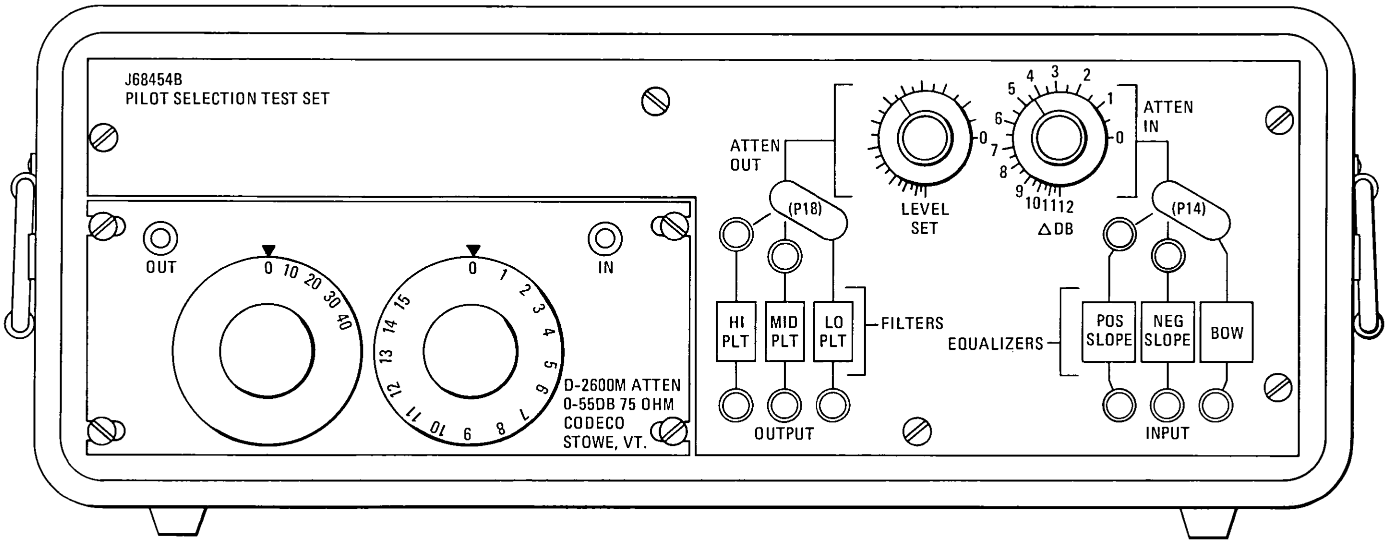


Fig. 6—Pilot selection test set.

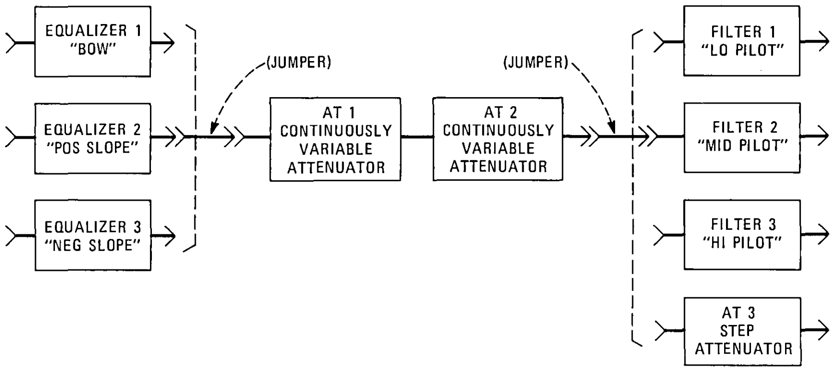


Fig. 7—Block diagram of the pilot selection test set.

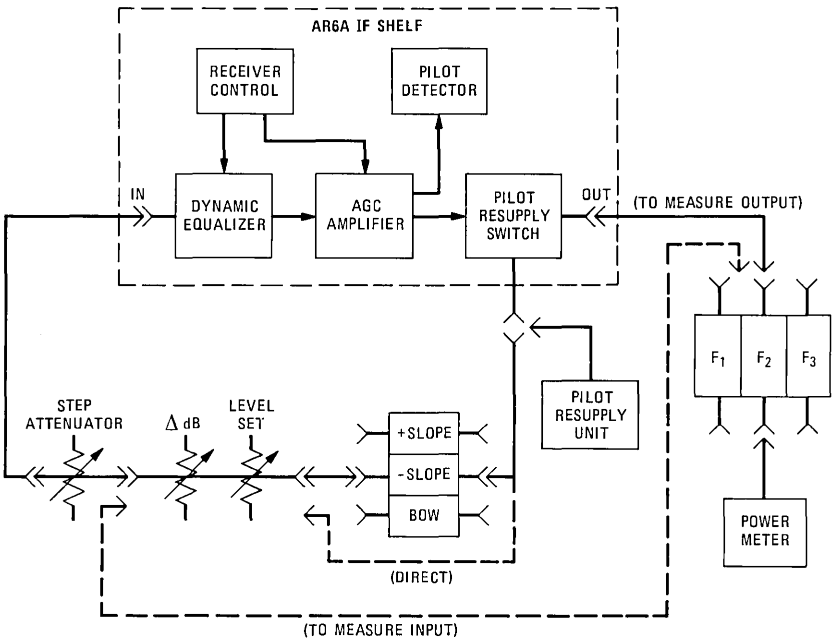


Fig. 8—Shelf-test functional diagram.

pilot source that is used for the pilot resupply mode of operation of the TR bay. The pilot resupply unit supplies the fixed amplitude pilot tones and the PSTS then modifies their amplitudes as required to test the AR6A bay (see Fig. 8).

The output from the pilot resupply unit contains all three pilot tones mixed together. The PSTS contains a narrowband monolithic crystal filter for each of the tones. These filters, when used in conjunction with a power meter, allow each tone to be measured individually. This arrangement is used to set the pilot reference power levels initially and then to observe the pilot levels after being acted upon by the AR6A receiver under test.

The PSTS checks the performance of the AR6A dynamic equalizer by first establishing equal pilot signal levels and then introducing known level differences in the pilots by the use of shape networks. Signal correction by the dynamic equalizer can then be observed.

The three shape networks are a positive slope, a negative slope, and a bow shape. All three shape networks attenuate the center frequency pilot 10 dB. The positive slope network attenuates the low-frequency pilot 18 dB and the high-frequency pilot 2 dB. The negative slope network attenuates the low-frequency pilot 2 dB and the high-frequency pilot 18 dB. Thus, the two networks can exercise the end pilots relative to the center pilot by ±8 dB. The bow network attenuates the center pilot frequency 10 dB and both the upper and lower pilot frequencies 2 dB. The resulting signal transmission through the network will be a bow with the end pilots 8 dB higher than the center pilot.

The PSTS checks the performance of the AR6A AGC amplifier by varying the amplitude of all three pilots together with a step attenuator while monitoring the AGC function. A precision step attenuator is provided to adjust signal levels over a 55-dB range in 1-dB steps. This step attenuator can also be used with the linearity test set if desired. The pilot resupply and space-diversity threshold levels are checked by varying the amplitude of all three pilots together with a continuously variable attenuator while watching for trip levels to occur.

Two continuously variable attenuators are provided. One is used to set the input power level accurately while the other is used to determine trip levels. The use of two variable attenuators allows the input power level adjustment to remain undisturbed while the second attenuator searches for trip levels. The original test set input power level can then easily be restored by turning the second attenuator back to zero. The use of continuously variable attenuators eliminates the transient effects on trip levels associated with step attenuators.

For convenience, hairpin patch plugs simplify the interconnection of the shape networks to the variable attenuator input. Also, a second hairpin plug interconnects the variable attenuator output to the pilot filters.

IV. CONCLUSION

The scanning IM test set has proven to be essential in the field for the initial transmitter predistorter linearity alignment and for the continued confirmation and maintenance of the high degree of linearity required by the AR6A Radio System.

The pilot selection test set is essential in the field to verify the correct functioning of the IF receiver circuits and to provide simulated fading for checking trip levels of the pilot resupply and space-diversity circuits.

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ACRONYMS AND ABBREVIATIONS

ACO	alarm cutoff
AGC	automatic gain control
AM	amplitude modulated
AR6A	amplitude modulation radio at 6 GHz, version A
BEF	band-elimination filter
BFE	beam-forming electrode
BPF	bandpass filter
BSP	Bell System Practices
BSR	Bell System reference
BSRF	Bell System reference frequency
BSRT	Bell System reference tone
BSRTU	Bell System reference transmitting unit
CMOS	complementary metal-oxide semiconductor
CRT	cathode ray tube
D/A	digital-to-analog
DAS	Data Acquisition System
DDD	direct distance dialing
DUT	device under test
EMI/RFI	electromagnetic interference/radio frequency interference
EQ	equalizer
FB	front-to-back ratio
FET	field-effect transistor
FSK	frequency shift-keyed
G/T	gain/temperature
HPF	high-pass filter
IF	intermediate frequency
IM	intermodulation
JFS	jumbogroup frequency supply
LC	inductor capacitor
LED	light-emitting diode
LIF	lower-band intermediate frequency
LO	local oscillator
LO	locally generated microwave carrier
LOA	law of addition
LPF	low-pass filter
LSB	lower sideband
MCSS	microwave carrier synchronization supply
MG	mastergroup
MGDF	mastergroup distributing frame
MGTB	mastergroup translator, type B
MMG	multimastergroup

MMGT–R	multimastergroup translator–radio
OMFS	office master frequency supply
PAL	phase alternation line
PFS	primary frequency supply
PLL	phase-locked loop
PM	phase modulated
PSTS	pilot selection test set
R-C	resistor-capacitor
RF	radio frequency
RFS	regional frequency supply
SC	suppressed carrier
SCOTS	Surveillance and Control of Transmission Systems
SDC	Shottky diode converters
SDS	Software Development System
SECAM	sequential with memory
SL	section loss
SPDT	single-pole double throw
SSB	single sideband
SSBAM	single-sideband amplitude modulation
STMS	selective transmission measuring set
TASC	telecommunication alarm surveillance and control
TDMA	time division multiple access
TL	transmission level
TR	transmitter-receiver
TSS–R	Transmission Surveillance System–Radio
TTL	transistor-transistor logic
TWT	traveling-wave tube
UIF	upper-band intermediate frequency
USB	upper sideband
VCO	voltage-controlled oscillator
VDC	varactor diode converters
XPD	cross-polarization discrimination

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