A New Wide-Band System Architecture for Mobile High Frequency Communication Networks

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Abstract—A new wide-band HF communication system architecture for a multiple, mobile terminal network has been proposed and analyzed. This architecture embodies the complete removal of high power switching and tuning and achieves control of unwanted radiation by the use of highly linear power amplifiers and adaptive interference cancellation methods. To provide maximum flexibility in system operation, provision is made for transmitting and receiving subsystem assets to be shared among users to the greatest degree possible. The transmitter is simply a power bank into which all exciters are connected, and each receiver has access to a selection of antennas through power dividers, diversity combiners, and adaptive array processors.

The implementation of this architecture will permit rapid and flexible frequency selection and channel configuration to accommodate the complex network relaying strategy that will be necessary in modern warfare environment.

Research is being conducted by the U.S. Navy on the technologies that require advancement if this architecture is to be brought to fruition for intratask-force communication. A system engineering analysis, substantial measurements and simulations of adaptive cancellation system performance, and development of critical system components have been conducted to validate the architectural concept.

INTRODUCTION

It is our purpose in this paper to describe a new radio frequency (RF) system architecture for high frequency (HF) mobile communication networks whose evolution in the recent few years has resulted from a combination of new communication needs and the availability of newly developed electronic technologies.

As a consequence of the obvious physical and electronic vulnerabilities of the satellite communication systems that are likely to be available in this century, the HF band has substantial promise for the conduct of military tactical communication. However, the technology used in the present military HF communication system is a relic of the 1950's that has languished until recently in the backwaters of our love affair with the satellite. Our present HF communication system architecture cannot provide the connectivity under stress that a tactical communication system must provide. Its failure stems from inflexibility in the high power RF hardware and in the RF distribution subsystems, which contain mechanically tuned and switched narrow-band components. It contains little provision for channel evaluation and link performance assessment or for fundamental frequency emission amplitude control to limit electromagnetic interference (EMI).

At the ranges of a few hundred kilometers involved in most naval intratask-force tactical communication missions, the surface wave mode of propagation is available most of the time. The RF system architecture we have developed anticipates use of this mode, with its relatively dispersion-free propagation and modest frequency dependence by comparison to skywave. We propose the use of a relaying concept to exploit this limited frequency dependence in combatting jamming by making our most vulnerable link as short as possible and forcing a jammer to attack this link at a frequency that is to his greatest disadvantage. Reliance on this relaying concept to overcome the challenges with which a modern tactical communication system must cope requires that a number of features be organic to the system architecture:
- rapid and flexible selection of transmitted frequency,
- compatibility with interference resistant waveforms,
- channel assessment techniques that permit link and network performance monitoring instantly and continuously,
- stringent amplitude control of RF emissions, and
- rapid on-line system reconfiguration and adaptability.

Our purpose in this paper is to describe the RF subsystems that will be necessary to accommodate such a concept.

The new RF architecture is fully wide-band and provides for rapidity of frequency change that is limited solely by synthesizer settling time. It is characterized by complete elimination of tuning in the RF distribution system and the complete elimination as well of high power switching. It exploits sharing of assets to the greatest feasible extent to permit essentially unlimited flexibility in numbers of available circuits and in the distribution of bandwidth and radiated power among these circuits. Of course, there is no such thing as a free lunch, and the flexibility gained by this schematism is purchased at the price of extraordinary measures to control EMI. There is, after all, nothing revolutionary about the concept of a wide-band RF architecture. The system known as ICS-3 (Integrated Communications System 3), which is in production for the British Navy, embodies a wide-band RF architecture [1], as does a proposed HF communication system for the Canadian Navy [2].

The intermodulation (IM) products and noise generated in all parts of the RF hardware must be removed before radiation or reduced sufficiently in the receiving subsystem if acceptable receiving sensitivity is to be preserved. None of the scrupulously designed narrow-band transmitting subsystem filters and multicauplers, and none of the preselectors and tuned multicauplers in the receiving subsystem that characterize our 1950's era solution to the EMI problem is available to us in this new architecture; we must seek new means of overcoming these unwanted radiations.

Subsequent sections of this paper will describe the major system components of our proposed wide-band RF architec-
ture. The greatest attention will be concentrated upon provision for EMI control, the principal subject matter of our research for several years. Concluding sections will present our attempt to evaluate the technical challenges that we believe are entailed in reduction to practice of this architecture.

Before we undertake description of the proposed system, however, we need some critical pieces of information to establish realistic goals for receiving subsystem sensitivity and for transmitting subsystem radiated interference. These parameters are of importance for two reasons.

1) The vulnerability of a receiver to internally generated IM products is directly related to the amount of fundamental signal power coupled to the receiver from colocated transmitters. We should strive to design a receiving subsystem whose internal noise is comparable to the noise floor established by the dominant environmental noise sources, and we should implement this design in a manner that minimizes mutual coupling with nearby transmitting antennas.

2) The radiation by a transmitter of IM products and other interference can itself establish an effective noise floor, and care must be taken to assure that this floor does not exceed otherwise dominant environmental sources. The radiated IM products ideally should not exceed those that are beyond the system designer’s control; in the case of naval communication systems, an example is IM products created by nonlinearities in the ship’s topside environment (the familiar “rusty bolt” effect).

We select for a specific application here the example of a naval combatant platform and use as goals for receiving and transmitting subsystem performance the noise and topside IM environments characteristic of such a platform. We have two sets of data to aid us in establishing these goals. As a result of HF noise measurements made by the National Bureau of Standards [3] and measurements of noise made themselves aboard four U.S. Navy ships, Gustafson and Chase [4] have defined the parameter quasi-minimum noise (QMN), which is intended to represent typical low-noise periods in low-noise environments (not including the Arctic, however). The levels of QMN spectral density at four frequencies are listed in Table I. Although true environmental noise occasionally drops below these levels they represent practical working data for system engineering purposes and have served in our analyses as performance goals for both receiving subsystem sensitivity and transmitting subsystem radiated broad-band noise coupled to the receiver.

Secondly, Salisbury [5] has measured the IM products generated in the topside environment of three U.S. Navy ships that were designed with deliberate attention to EMI control by use of nonmetallic topside fixtures, where feasible, and careful attention to bonding of metallic structures. These data were taken with standard 1 kW U.S. Navy transmitters of late 1960’s and early 1970’s vintage, using broad-band transmitting antennas that do not depart significantly in configuration from those we envision in our proposed architecture. Third-order IM products measured in Salisbury’s program are shown in Fig. 1 as a scatter plot of data from several experiments in comparison to QMN in a 3 kHz bandwidth. These data have been used in our analyses as a goal for transmitter-generated maximum third-order IM products.

### GENERAL FEATURES OF THE NEW WIDE-BAND HF ARCHITECTURE

The new architecture possesses four prominent features.

- The full HF spectrum is available to all users at all times.
- There is no RF tuning between the exciters and the transmitting antennas or between the receiving antennas and the receivers.
- Rapidity of frequency change is limited by synthesizer settling time and the settling time of adaptive circuitry (and is estimated to be of the order of a millisecond).
- Minimum separation between transmitting frequencies and between transmitting and receiving frequencies is limited only by synthesizer phase noise (and is estimated to be about 2 percent).

Fig. 2 is a conceptual diagram of the transmitting subsystem in the new architecture. Signals from several types of exciters are combined in a single exciter combiner. Narrow-band signals consist of a conventional voice (3 kHz bandwidth) and radio teletype (16 channels of 100 Hz bandwidth, each in 3 kHz total bandwidth) services. Broad-band signals are envisioned as occupying up to 100 kHz instantaneous bandwidth and may be frequency hopped at a rate of a few hundred hops per second over any hopping bandwidth. The channel evaluation signal probes the entire HF spectrum, or the parts of it whose use is anticipated, in a format that minimizes interference with on-line users, but characterizes candidate communication channels in near real time.

The exciter combiner feeds an input hybrid to a modular bank of several wide-band power amplifiers. We expect that such amplifiers of about 1 kW output power rating each will constitute a typical power bank. An output hybrid combines the signals from these power amplifiers and feeds two wide-band transmitting antennas.

Several reference signals are tapped as shown for use in
EMI reduction. We have identified an "EMI control subsystem" in Fig. 2 solely for convenience in describing our architectural concept; likely implementations of the architecture may well find EMI control circuits distributed individually within the other subsystem components. The EMI control scheme comprises a number of adaptive elements. Some of these circuits will make use of fundamentals sensed in the transmitting subsystem to effect cancellation of fundamentals coupled into the receiving subsystem via the local propagation channel. Others will be used for adaptive interference control within the transmitting subsystem itself by feedforward techniques.

Fig. 3 is a conceptual diagram of the receiving subsystem in the new architecture. The wide-band receiving antennas are physically and electrically small, and each is connected to an RF power divider through an individual wideband, high dynamic range, low-noise distribution amplifier. Power divider outputs are available for any desired receiving function in any desired combination. Thus, this architecture is compatible with the extensive employment of adaptive arraying and diversity combining for both narrow-band and broad-band signals and for the channel evaluation function.

The absence of all high power switching and tuning in this new architecture suits it to a simplified automatic control scheme that is the key to implementation of the five features listed in the Introduction to this paper. All tuning is done in the synthesizers and adaptive control circuitry; all power control is accomplished in the exciters; there is no patching or switching. The availability of the full HF spectrum to all users makes possible unlimited operational flexibility. Any combination of broad-band and narrow-band signals may be radiated simultaneously. Narrow-band signals may even be emitted within the hopping bandwidth of frequency hopping broad-band signals. Channel evaluation functions, both sounding and interference monitoring or occupancy scanning, are interleaved with communication functions and use the same RF assets. This process can be accomplished by a single nodal controller which can also: 1) monitor circuit performance, 2) ration assets to favor high priority users, 3) minimize EMI by employing frequency and power control, and 4) respond to changes in traffic demand and priority by reconfiguring the entire system with ease. This feature permits sudden and frequent changes in network connectivity when necessary to cope with interference, jamming, propagation outages, losses under hostilities, and other aspects of network dynamics.

The use of shared assets in both the transmitting and receiving subsystems provides the necessary flexibility to meet reasonable goals for system availability and allows graceful degradation. For example, Fig. 4 shows the effect upon total radiated power owing to the failure of 1 kW power amplifier modules in a 16 amplifier power bank. Four simultaneous amplifier failures lead to a 2.5 dB loss per signal, resulting in 9000 W total available output power. Thus, the system can sustain modest losses without catastrophic failure.

The use of miniature receiving antennas has similar consequences for the receiving system. Such antennas, when combined with distribution amplifiers and receivers of sufficiently low internal noise to establish a subsystem noise floor equal to QMN, can be made small enough to allow installation at points on the platform that are far removed from the transmitting antennas. This circumstance aids in the reduction of EMI coupling between the two subsystems. These antennas also can be made sufficiently unobtrusive so that many of them can be installed on a mobile platform of limited size, thus permitting the exercise of adaptive arraying and diversity combining techniques and leading to asset sharing in the receiving system that provides inherent graceful degradation.

Because of the absence of tuning, of course, the entire system must display very high linearity: in the transmitting subsystem to reduce radiated IM products, and in the receiving subsystem to reduce internally generated IM products. Broadband noise radiated by the transmitting subsystem must also be kept below strict limits. The key system components from the standpoint of noise and IM performance are: 1) synthesizers, 2) RF power amplifiers, 3) distribution amplifiers, 4) receivers, and 5) interference cancellers.

The EMI control scheme we have suggested in Figs. 2 and 3
Fig. 4. Degradation in output of power bank consisting of 16 1 kW amplifiers as amplifier failures accumulate.

employs adaptive and passive feedforward distortion control in the power bank. The scheme uses adaptive interference cancellation (AIC) before radiation of noise and intermodulation products that are generated in the bandwidths of the receivers in use. AIC techniques also are used for the cancellation of transmitted fundamentals coupled into the receiving subsystem in order to control IM products generated in the receiving subsystem.

Should the implementation of this architecture be successful, several advantages beyond those of operational flexibility will be realized. For example, system growth potential will be limited only by spectrum availability, platform prime power, and platform real estate; exciters, power amplifiers, receivers, and receiving antennas can be added at will within those limits.

Total system availability will be enhanced by the absence of high power electromechanical devices and by the removal of automatic circuitry used in many cases for the control of these devices.

The inherent modularity of the asset sharing concept will permit the system to be repaired essentially on-line by isolating failed components for replacement while other assets temporarily assume a slightly heavier burden. Susceptibility to system failure will be substantially reduced by the elimination of single-point failure nodes through asset sharing.

The skill levels necessary for system operation will be substantially reduced inasmuch as all tuning of highpower devices will be eliminated. The compatibility of this new architecture with microprocessor control by a single nodal manager will allow the incorporation of built-in testing, thus eliminating the need to troubleshoot an extensive, dedicated chain of equipment in the event of component failure. There will be fewer functions to control, and the simple RF distribution scheme will make it possible for a single nodal controller to exercise effective system control.

The exercise of nodal control by a single manager with continuous channel availability and system performance information will remove the need for operator guesswork in setting up communication circuits.

**TECHNICAL CHALLENGES POSED BY THE NEW ARCHITECTURAL CONCEPT**

The performance goals to which we have referred in the Introduction to this paper require that: 1) the aggregate of all IM products generated within the total system, referred to a common point in the receiving subsystem, must not be greater than the corresponding level of platform generated IM products; and 2) the aggregate of all noise generated within the total system, referred to a common point in the receiving subsystem, must be no greater than QMN. Our design approach for achieving this performance requires: 1) development of highly linear broad-band power amplifiers, 2) the use of miniature receiving antennas and low noise receiving subsystem electronics to facilitate transmitting/receiving subsystem decoupling, and 3) the use of AIC techniques to remove unwanted IM products and noise before radiation from the transmitting subsystem and unwanted fundamentals in the receiving subsystem.

We have performed an analysis of the noise, harmonics, and IM distortion characteristics of a state-of-the-art (SOA) wide-band HF communication system employing a 1 kW broad-band HF power bank similar to that which is used in ICS-3 and broadcasting multiple-tone fundamentals that simulate a signal structure akin to military signaling practice. The analysis was based on measurements of the power bank provided by the Admiralty Surface Weapons Establishment (ASWE). With the power bank delivering four 250 W tones the distortion performance is reported in Table II.

For our analysis, we estimated distortion power density for a case of 64-tone emission distributed equally among four fundamentals of 3 kHz bandwidth and 250 W per fundamental. Our principal interest is the occupied bandwidth of harmonic and IM distortion product power and the levels of IM products relative to the topside generated level. For this calculation, we selected fundamentals of 3272, 4407, 6175, and 15 190 kHz. If each fundamental is then assumed to be modulated by 16 tones distributed over a 3 kHz bandwidth, a band of distortion power will be centered upon each of 2271 distortion products and will occupy a bandwidth equal to the fundamental bandwidth multiplied by the order of the distortion product. The amplitude of each distortion-product order is computed based on the intermodulation values of Table II. The resultant power is assumed to be uniformly distributed over the occupied bandwidth of the unique distortion-product families. The results have not been adjusted upward to account for the rise in total distortion power predicted by theory [6] for an increase in numbers of tones from 4 to 64. The error in this assumption is considered to be relatively small, at least for lower distortion-product orders; it is approximately 3 dB for fifth order. Fig. 5 shows the bandwidths occupied by distortion products through eleventh order in the case we have treated, related to fundamental level (left-hand scale) and in absolute power for 1 kW total radiated power (right-hand scale).
The eleventh-order products of the output of the transmitting subsystem occupy about 65 percent of the HF band at a level as large as $-65\,\text{dBW}$ per 3 kHz bandwidth (or $-100\,\text{dBW/Hz}$ power spectral density).

The effect of this distortion power on a collocated receiving subsystem can be estimated. If, for example, the receiving antenna is located 30 m from the transmitting antenna, with both antennas sited over perfect ground containing no other reflecting surfaces, Fig. 6 shows the level of distortion products at the receiving antenna to eleventh order relative to QMN. The eleventh-order products, which have been shown to occupy most of the HF band, are more than 40 dB above QMN and thus form an effective system noise floor.

Radiated broad-band noise can be given a similar treatment, although its impact upon the receiving subsystem requires some assumptions regarding the design details of that subsystem. For the 1 kW SOA power amplifier whose distortion products we analyzed, broad-band noise is emitted at a level of $-100$ to $-105\,\text{dBW}$ in a 3 kHz bandwidth ($-134$ to $-139\,\text{dBW/Hz}$) separated by 2.5 percent from an operating frequency. Its principal source is the exciter synthesizer. Assuming that this power amplifier is matched to an 11 m whip and is separated by 30 m from a receiving subsystem whose receiving element is a 2 m whip, Fig. 7 shows the ratio of received broad-band noise power to QMN over the HF band. The shape of the response curve is related to the radiating characteristics of the 11 m whip. Broad-band noise exceeds QMN by more than 10 dB over this band.

Lower order distortion products are of concern by comparison to those generated in the topside environment. For the power amplifier treated above, third-order intermodulation products are 50 dB below fundamental levels for a two-tone case with 500 W per tone, and $39-69\,\text{dB}$ below fundamental levels for a four-tone case with 250 W per tone. For the simple platform scenario treated above, the level of third-order distortion at the receiving antenna for two-tone emission exceeds topside generated IM products by an average of 50 dB over the HF band.

To summarize, an SOA broad-band power amplifier fails to meet our design goals by amounts that vary from 44 to 55 dB, for eleventh-order distortion products that form an effective noise floor, to as much as 60 dB, for low-order distortion products that must be compared to those generated in the topside environment. Radiated broad-band noise falls 10-20 dB short of permitting our goal of QMN for receiving subsystem sensitivity to be met.

In contending with IM products generated within the receiving subsystem, high dynamic range electronics will be required. However, decoupling of the receiving and transmitting subsystem by careful attention to antenna siting and by the use of electrically small receiving antennas has the benefit of providing a multiplying effect when related to IM products generated in the receiver: the level of third-order products produced in the receiver varies as the cube of received fundamental signal power. This circumstance makes it attractive to achieve as much spatial decoupling as possible. The remainder of the receiving subsystem IM performance must be achieved.
by the use of high dynamic range components or by the adaptive schemes we have suggested.

A miniature receiving antenna for HF was designed and tested several years ago [7]. This design employed a 1 m monopole with a 1 m top-loading disk and, when combined with SOA electronics, displayed a noise floor below QMN over the entire HF band.

A tradeoff is possible between the noise performance of the electronics in the receiving subsystem and their distortion performance: improvements in component noise performance will allow greater decoupling to be effected between the transmitting and receiving subsystems, thus easing requirements on subsystem IM performance. For the platform scenarios we have analyzed, Fig. 8 relates the required third-order output intercept point [8] (OP13P) of the receiving antenna distribution amplifier to the aggregate noise factor of subsequent elements of the receiving subsystem. As the noise factor decreases, the OP13P requirement eases. Assuming that an aggregate noise factor of 6 dB is attainable in these subsystem components, the antenna amplifier will have to exhibit an OP13P of about 43 dBW. The corresponding IP13P requirement of the associated receivers would need to be 33 dBW (allowing for 10 dB loss in the power divider). The current state-of-the-art for receiver IP13P performance is considered to be about 3 dBW (noise figure of 10 dB) and is a limitation imposed by the mixer. Either a vast improvement in mixer performance or reduction of fundamental signal amplitudes by about 30 dB is required to permit use of current receivers. The use of AIC for reduction of fundamental signal amplitude is feasible for this application and if introduced prior to the distribution amplifier reduces the OP13P requirement of the amplifier and subsequent components in the receiving subsystem by the amount of cancellation actually attained.

IMPLEMENTATION ALTERNATIVES FOR THE NEW ARCHITECTURE

Given the limitations of SOA technology, how do we either advance that technology or compensate for its deficiencies so that our proposed architecture can be implemented? Three avenues are open to us: 1) improvements in transmitting subsystem linearity and noise performance, 2) improvements in receiving subsystem linearity and noise performance, and 3) extensive employment of interference cancellation techniques.

To assure that communication system noise performance will be dominated by QMN, noise power originating in the transmitting subsystem should be no greater than one-tenth of QMN at the receiving subsystem, thus permitting the allowable noise introduced by the receiving subsystem to be essentially equal to the value of QMN when referred to a common point in the receiving subsystem. For the platform scenario we have analyzed, the maximum permissible equivalent transmitting subsystem radiated broad-band noise level is \(-167\) dBW/Hz. We believe this performance can be approached, but it will require the imposition of a minimum separation between transmitting and receiving frequencies of nominally 2 percent or 100 kHz (whichever is greater) and improvements in exciter noise performance.

We have indicated above that realistic receiving antenna and distribution amplifier combinations can provide QMN sensitivity. Further improvements in noise performance of all components in the receiving subsystem can be traded off to permit a less stringent OP13P requirement. However, the receiving subsystem probably will require cancellation of fundamental signals.

Fundamental signal AIC (FSAIC) will be used for this purpose and will employ a reference signal tapped at the exciter outputs to control a cancellation signal that will be injected at the input to the distribution amplifier. This approach avoids the need for the cancellation processing circuit to handle more than one fundamental and thus eases the distortion performance requirements of the processing circuits. The application of AIC to naval communication systems is described in [9]. The objective of an FSAIC is to effect cancellation of the portion of the transmitted signal that is coupled into the receiving subsystem by processing a sample of the transmitted signal to be equal in amplitude but 180° out of phase with the interference at the point of cancellation in the receiving subsystem. Because of the numerous reflecting surfaces on a military platform, the local channel is highly dispersive and characterized by a complex phase and amplitude versus frequency response. The precision with which the processed sample matches the phase and amplitude characteristics of the interference channel over the bandwidth of interest determines AIC performance. Implementation of the FSAIC will employ a feedback loop that will sense changes in the propagation channel from transmitter to collocated receiver by means of an (interference) error signal at the distribution amplifier output. This loop will control the proportion of processed transmitted signal power injected for cancellation.

To verify general feasibility of the FSAIC concept in a naval context, a series of measurements was performed aboard the U.S.S. MILLER (FF1091) and U.S.S. AINSWORTH (FF1090). The purpose of these measurements was to: 1) determine the amplitude and phase characteristics of the local channel existing between shipboard transmitting and receiving antennas, 2) use these measurements in a computer simulation to assess
AIC performance, and 3) determine experimentally the response characteristics of a (manually adjusted) single tap AIC system for the shipboard channel.

For these measurements, an 11 m whip was used as the transmitting antenna and an electrically small antenna, consisting of a 2 m whip transformer coupled to an amplifier, was employed as the receiving subsystem. Fig. 9 shows the amplitude and phase (relative to an arbitrary reference) characteristics of the channel existing between the two antennas in the region of 8.0 MHz, referred to the input to the transmission line driving the 11 m whip. The amplitude results of Fig. 9 include the gain of the amplifier used with the electrically small antenna and show the signal level delivered to the receiver with respect to the power available from the source generator applied to the transmitting whip, including losses due to impedance mismatch between the source generator and the transmitting whip. The two traces in each graph were taken two min apart and show variations that may be typical in a realistic environment. Amplitude varies by about 1 dB and phase by about $15^\circ$ in a 200 kHz bandwidth centered at about 8.0 MHz.

Fig. 10 shows the amplitude performance of a single tap, manually adjusted FSAIC circuit operating at a center frequency of 8.04 MHz in the channel of Fig. 9. Amplitude response without the AIC is reproduced in Fig. 10, as well. The two traces of AIC response illustrate further the variations in the amplitude and phase properties of the local channel that prevent a nonadaptive interference cancellation system from maintaining a constant level of performance. The bandwidth over which cancellation was effected at a level of suppression of 30 dB varied from 22 to 40 kHz.

The basic channel data of Fig. 9 were also used in a computer simulation of an AIC system to assess performance. Fig. 11 shows the results of this simulation at a center frequency of 8.04 MHz. The bandwidth for 30 dB of cancellation is about 27 kHz and is gratifyingly near the cancellation bandwidth achieved experimentally with a single tap circuit. In simulations at many frequencies over the full 2-30 MHz range, the RF bandwidth over which 30 dB or more of cancellation can be attained averages about 30 and 80 kHz, respectively, for single and dual tap arrangements. We believe that cancellation levels of 30 dB and greater can be achieved over bandwidths between 10 and 100 kHz, depending upon the nature of the local channel, with modest numbers of taps. This performance will permit attainment of IM performance requirements in the receiving subsystem.

Previous designs of HF AIC systems [10] have not been intended to function in the multiple signal environment characteristic of military tactical platforms, and there is some risk that if the FSAIC is required to handle a multiplicity of large signals, distortion generated within the cancellation circuitry could limit AIC performance. Use of a separate processor for each exciter in our approach minimizes the requirement for processor distortion performance, but does not eliminate the IM generation problem entirely because the combiner that collects the processed FSAIC signals does not provide perfect isolation between signals. It is thus important to minimize the amplitude of the processed signals handled by the AIC.

The risk in development of the FSAIC also is dependent on receiving subsystem noise performance. Fig. 12, computed for a shipboard channel similar to those we have treated above, shows the variation with frequency of power to be cancelled by the FSAIC when the noise factor of the cascaded distribution amplifier, power divider, and receiver is 3.5 dB. The power to be cancelled is maximum near 16 MHz and is about $-14$ dBW. Inasmuch as the success of interference cancellation techniques has been limited primarily by IM generated within the cancellation circuitry [10], the ability of our AIC circuitry to handle powers of that magnitude represents a substantial technical challenge. The magnitude of third-order topside generated IM products in a 3 kHz bandwidth is roughly 40 dB greater than QMN (3 kHz bandwidth) from Fig. 1. The corresponding amplitude at the input of the distribution amplifier is about $-125$ dBW for the receiving subsystem we have used in
our calculations as described above. We wish to ensure that the
distortion generated by the FSAIC processor is at least 10 dB
less than the magnitude of topside sources at the worst case
frequency of 16 MHz. This choice imposes an OIP^3P require-
ment of about 50 dBW on the FSAIC processor and requires
the FSAIC processor to develop cancellation signals up to -11
dBw (allowing 3 dB loss in the FSAIC combiner hybrid) and
to be free of residual third-order distortion to a level of -132
dBW, an intermodulation ratio of about 121 dB.

The major broad-band components likely to be employed
in an implementation of an AIC consist of quadrature hybrids,
0°-180° hybrids, directional couplers, power splitters, elec-
tronically adjustable attenuators, delay lines, mixers, low-pass
filters, and amplifiers. Because of these stringent self-generated
distortion requirements of an AIC system, representative sam-
ple of some of the components were purchased for evalua-
tion purposes. Fig. 13 shows as an example, the spectral re-
response characteristics of an off-the-shelf small signal hybrid
combiner transformer when handling two fundamental signals
at 2.1 and 3.0 MHz, respectively, with amplitudes of -17 dBW
(1 V) each. The major responses at 2.1 and 3.0 MHz in Fig. 13,
amatically -17 dBW, are the fundamentals reduced in amplitude
by band-stop filters preceding the spectrum analyzer. The
magnitude of the strongest distortion component is about -58
dBm (-88 dBW), yielding an intermodulation ratio of only
71 dB. The corresponding OIP^3P is about 18.5 dBW. This per-
formance is considered typical for small signal devices of this
type, and although adequate for some applications, is intoler-
able for an FSAIC system. The primary source of distortion
in each of the components examined is the ferrite core material.
The data from this and other tests show a definite trend to-
dward reduced distortion amplitudes for components rated at
high power levels (50 W or more), probably because of their
use of larger volumes of core material with consequent lower
flux densities. Accordingly, it is believed that the required dis-
tortion performance can be achieved with directional couplers,
various types of hybrids and similar components, but attention
must be paid in their development to achieving superior IM
performance.

Distortion products generated in the transmitting power
amplifiers must be reduced before radiation or cancelled in the
receiving subsystem because of their potential desensitization
of the collocated receiving subsystem. Typical two-tone third-
order distortion product levels for 1 kW wide-band solid-state
power amplifiers are (conservatively) about 30 dB below funda-
mentals; the SOA power bank of the ICS-3 system is capable
of attaining approximately -50 dB. For transmitting subsys-
tem distortion products to fall 10 dB below topside generated
distortion (referred to the receiving subsystem) third-order
products must fall 120 dB below fundamentals for typical
shipboard antenna siting arrangements. The employment of
negative feedback in linear amplifier design, Class A operation
and use of large numbers of the most linear solid-state devices
available are expected to yield -55 dB performance, a modest
improvement with respect to ICS-3.

This performance will be extended to -70 dB by means of
a passive feedforward circuit arrangement [11], a simplified
version of which is shown in Fig. 14. The directional coupler
designated DCA in Fig. 14 obtains a sample of the input funda-
mentals, which is delivered to equalizer 1. The output signal
from equalizer 1 is delivered to summer Σ_A. Also delivered
to summer Σ_A is a sample of the signals existing in the output
of power amplifier block A_A that is obtained by means of di-
rectional coupler DCB. If the two samples of fundamental
energy arriving at Σ_A are equal in amplitude, time coincident,
and 180° out of phase, they will be cancelled. If cancellation
of fundamentals is perfect, the signals at the input to error
amplifier A_2 will consist only of noise and distortion signals
that were generated in the power amplifier modules and
hybrids. Error amplifier A_2 amplifies the sample of noise and
distortion generated within the power bank. The amplified
sample of noise and distortion is delivered to directional
coupler DCC, which follows equalizer 2 in the main output

Fig. 12. Power required to be cancelled at input of wide-band low
noise distribution amplifier; used with electrically small antenna 30
m from an 11 m whip radiating two 1 kW tones.

Fig. 13. Distortion performance of off-the-shelf small signal hybrid
combiner transformer. Fundamental signals applied at 2.1 and 3.0
MHz at +13 dBm (1.0 V) each.
path. The gain of amplifier $A_2$ is exactly equal to the sum of the losses introduced by, directional couplers DCB, DCC, summer $\Sigma_A$ and equalizer 2. The function of equalizer 2 is to insert a delay in the main path so that time coincidence is achieved with signals delivered by $A_2$.

Feedforward distortion cancelling techniques are limited in performance by the phase and amplitude matching that can be maintained in the cancelling loop. For example, to achieve cancellation of 20 dB, the amplitude must be matched to within approximately 0.5 dB and the phase to within approximately 4°, performance which is well within the capability of the state-of-the-art [12]. Noise and distortion originating in the error amplifier are not cancelled by this method.

The performance of the feedforward loops can also be degraded by power reflected from the transmitting antenna and delivered to the output of $A_2$ through DCC. The intermodulation products generated as a result of this reverse power must be less than the residual distortion of the hybrid combined basic amplifiers after cancellation by the feedforward loops.

The resultant estimated distortion performance (−70 dB) using the passive feedforward distortion control is still considerably short of our objective of −120 dB. AIC is applicable for reduction of these residual IM products generated in the transmitting subsystem. However, investigations conducted at NRL show that the complexity and resultant cost of providing full 2–30 MHz spectrum coverage for reduction of radiated IM products may be too high, owing to dispersion in the local channel. Because only IM products at frequencies being used for reception represent potential local interference, this second AIC need suppress only IM products appearing in the IF bandwidth of the receivers. We designate this second cancellation process IFAIC, and it can be accomplished either in the receiving subsystem or in the transmitting subsystem.

If IFAIC is implemented in the transmitting subsystem, the “channel” that must be tracked is expected to be much less complex than the local channel existing between transmitting and receiving antennas, and the IFAIC processor complexity can then be minimized. However, the amplitude of processed IFAIC signal required is much lower if the cancellation signal is injected into the receiving subsystem. Furthermore, noise introduced by the IFAIC is expected to be of less importance if the IFAIC is implemented in the receiving subsystem.

Selection of the proper IFAIC implementation must take into account a further circumstance. Unless transmitting subsystem generated IM products are suppressed before radiation takes place, these products may interfere with communication on nearby platforms. Fig. 15 shows the range for surface wave propagation over seawater at which third-order IM products from a 1 kW SOA amplifier are approximately equal to QMN for four fundamental signals of 250 W per signal. The potential interference range varies from about 241 km, at 2 MHz to 72 km at 17 MHz for a receiver bandwidth of 3 kHz. The proportion of IFAIC that should be achieved in the transmitting subsystem should consider not only radiated IM product effects on one’s own platform, but their effects also on platforms nearby.

Fig. 16 describes the IFAIC approach we are pursuing for the transmitting subsystem. The IFAIC consists of an AIC loop (Loop 2 in Fig. 16) that operates to minimize distortion products in the output of the power bank at frequencies being used for reception. Two auxiliary AIC loops are required in order to permit the main AIC loop to function. The function of the first auxiliary loop (Loop 1 in Fig. 16) is to cancel the fundamental signal from a sample of the power bank output and thus obtain a replica of the total distortion and noise introduced by the power bank.

A sample of the fundamental signal from each exciter is processed by a “weighting assembly” $W_1$. The processed samples are combined in the output of summer $\Sigma_1$ and delivered to directional coupler DCB. In order for cancellation to occur, the amplitude of fundamentals at the output of $\Sigma_1$ must be equal to and 180° out of phase with the sample obtained from the power bank output and launched toward amplifier $A_1$ by means of directional couplers DCA and DCB. The cancellation of fundamental components is controlled by an error signal developed in the output of $A_1$ for control of the $W_1$ assembly by Loop 3.

Loop 2 processes the sample of distortion that happens to appear in the bandwidth of the frequencies being used for reception. The processing is channelized and performed independently for each of eight receiver frequencies in the example shown in Fig. 16. Channelization is accomplished by the eight way splitter $\Sigma_2$, each output of which is associated with an IF weight block in which discrete distortion samples are upconverted to an IF frequency, filtered, amplified, weighted, and then down-converted to the original frequency. The output from each IF weight is then separately amplified by $A_2$ and delivered to the output combiner $\Sigma_3$. $\Sigma_3$ combines the processed distortion samples and delivers them to directional coupler DCD for injection into the power bank output after the equalizer.

Control of the cancellation process of Loop 2 is accomplished by extraction of an error signal (not shown in Fig. 16).
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Fig. 16. Adaptive feedforward distortion control (IFAIC) for reducing IM in output of power bank at frequencies used for local reception.

After directional coupler DCC for use in a second auxiliary AIC loop to cancel fundamental components appearing in the error signal path. Attainment of the performance goal appears feasible, but some risks exist in development of weights that have requisite power handling, distortion, and settling time characteristics.

While the preceding discussion has been directed toward suppression of third-order products, our goal is to suppress all distortion within the bandwidths of the local receivers to a level 10 dB less than topside generated products. A database does not exist for high-order topside products; however, we have arbitrarily assumed second- and third-order topside generated products to be equal and subsequent higher orders to be less in the amount of 10 dB per order.

The FSAIC and IFAIC employ reference signals from the exciters and, hence, are not capable of rejecting noise and distortion originating in them. Broad-band noise from the exciters is the dominant system noise source. The noise power emitted within AIC bandwidth by the exciter must be at least 190 dB below reference signal power. Attainment of required exciter noise performance together with requisite settling time, bandwidth, and programmability features is considered to be the most difficult challenge in the implementation of the wideband architecture.

CONCLUSION

The goal of achieving sufficient HF communication system flexibility and responsiveness to permit high connectivity and survivability in a dynamic network under attack by modern military means requires the use of a sophisticated relaying strategy. The RF system must permit rapid network reconfiguration under stress and must be designed to accommodate a wide variety of signaling waveforms and a large number of circuits. Careful attention must be given to continuous communication system performance assessment, including evaluation of both noise and propagation in the channel and identification of alternative channels. Antenna assets must be used effectively in arraying and diversity combining schemes to help optimize system effectiveness. Power control must be exercised both for improving link performance and for combating the local EMI environment. We have described in this paper a wide-band HF system architecture that can meet this goal.

The new architecture eliminates all high power switching and tuning and makes available the entire HF spectrum to all users. It exploits extensive sharing of transmitting and receiving subsystem assets to permit the maximum possible flexibility in system operation; it possesses an inherent graceful degradation feature. EMI control is achieved by adaptive interference cancellation techniques.

We have identified the technologies in which advancement must be made if implementation of the new HF system architecture is to be realized. The principal research and development areas are as follows:

- linear wide-band power amplifiers,
- low noise, high dynamic range RF distribution amplifiers,
- techniques for adaptive control of noise and IM products,
- low intermodulation-level electronic components, and
- low noise synthesizers.

A painstaking combination of component development and system engineering analysis will be required to achieve the balance of performance parameters, among many interactive system components, that can bring this concept to fruition. It is our belief that technology under development or readily within grasp will make a fully wide-band system architecture achievable within a very few years provided adequate attention is given to these research issues, most of which are currently under pursuit by the U.S. Navy.

The product of this effort will be an HF communication system whose flexibility will permit network connectivity and survivability under any realistic anticipated stress that far exceed both present capabilities and the capabilities of alternatives in other frequency bands.

ACKNOWLEDGMENT

The authors thank the personnel of ASWE, Portsmouth, England, for the open exchange of technical information concerning the ICS-3 system. We also thank R. M. Bauman of NRL for many contributions to the architecture, particularly in the area of adaptive EMI control techniques. The authors are grateful to D. C. Andrews of NRL for the development and implementation of the computer simulation program to predict FSAIC performance and D. Q. Arnesen of NRL for measurements of distortion performance of ferrite components. We also acknowledge the assistance of Zeger-Abrams, Inc., for analysis efforts of various AIC systems accomplished for NRL under Contract N00173-78-C-0238.

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