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A Pulsed Pseudo-Noise VHF Radio Set*

Abstract: The design of a radio set using pulsed pseudo-noise signals is described. The unit uses pulse position modulation and matched-filter detection. The use of pulse compression techniques permits the radio to achieve good signal detectability; it also permits privacy for each of several sets operating in a common channel. Characteristics of the 280 Mc/sec rf system are given and the call-up logic for achieving telephone-like operation is described. Some aspects of the digital matched filter design are discussed. Analysis of theoretical operation shows only a small performance difference between digital and linear matched filters.

Introduction

An experimental, portable VHF transmitter-receiver has been designed and built for co-channel operation. That is, several transmitter-receiver links can operate simultaneously in the same geographical location and in the same frequency channel, achieving telephone-like operation without the aid of a central switchboard. In this co-channel application, i.e., the so-called RADA (Random Access Discrete Addressing) concept,[1] communication is accomplished via a waveform referred to as "pseudo-noise" (PN) because of its noiselike properties. Pseudo-noise techniques are described in detail in a recent book by Golomb, et al.[2] Information describing the theory of random access, pseudo-noise communications systems may be found in a paper by Blasbalg[3] in this issue.

The envelope of the particular pseudo-noise signal used with this radio is a pulse 25.2 μsec long. It conveys audio information through conventional pulse-position modulation. The pulse itself is made up of 63 binary digits, or "chips," each lasting 0.4 μsec. The sequence of chips is used to modulate a 280 Mc/sec carrier signal. This is equivalent to double-sideband, suppressed-carrier modulation and is sometimes referred to as phase-reversal keying. (We prefer the term "chip" to "bit" in this discussion since each PN digit does not convey a bit of information in the classical sense of Shannon.[4])

The 63-chip sequence in the transmitted signal corresponds to the unique coding of a digital matched filter in the receiver set that the transmitting operator wishes to contact. To a receiver whose filter is not matched to the binary digit sequence being transmitted, the signal appears as noise.

When the PN pulse has been read completely into the digital matched filter of the intended receiver, an output pulse occurs. This output pulse has a width of one chip time, 0.4 μsec. The pulse compression thus obtained permits the radio to achieve good signal detectability; the PN pulse allows many radio sets to operate in the same frequency band with very little crosstalk.

The use of pulsed pseudo-noise signals is a primary difference between this and previous pseudo-noise radio units.[5] With a duty ratio of about 20%, the set overcomes some of the disadvantages of continuous pseudo-noise in a co-channel environment. For example, in satellite communications a big station-small station mix can be handled more effectively since interfering pulses from the higher power station will often fall between the lower power pulses without disturbing them. Moreover, pulsed pseudo-noise is readily adaptable to time-division multiplexing.

Analysis and simulation have shown that a network using the radio equipment described here can provide, ideally, an audio signal-to-noise ratio of 13 dB with approximately 13 equal-power talkers per megacycle. (A signal-to-noise ratio of 13 dB is equivalent to a digital matched filter error rate of about 0.24 for a system using 32 signal quantizing levels.) This performance figure is derived from the simulation by Van Blerkom, et al.[6] Thus the system appears able to support about 33 simultaneous equal-power talkers in a channel of 2.5 Mc/sec while providing an audible signal. The performance of the

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system with many simultaneous users has not yet been verified experimentally, however.

Description of the radio set

The experimental radio set is battery operated and weighs about 35 lbs; it has external dimensions of $16 \times 20 \times 9$ in. Since it was designed only for experimental purposes, no serious attempt at compact packaging was made—discrete components were used throughout. The system thresholds were made adjustable to facilitate laboratory experiments. In most applications of the device, the thresholds would be fixed, and the only operator controls would be an address selector and a volume control, as in ordinary telephony.

To use the set, the operator enters the address of the party he wishes to call into the transmitter's address register via a set of thumbwheel switches. He then pushes a call button which sounds a buzzer on the called set. Conversation can then begin.

A block diagram of the radio set appears in Fig. 1. The microphone of the handset is wired directly to a conventional pulse position modulator (PPM). At the instant a pulse is emitted from the PPM, it triggers the pseudo-noise generator (PNG), which then emits a 63-chip sequence. The particular sequence is determined by the contents of the transmitter's address register. During conversation the audio signal is sampled 8,000 times per second, allowing a 125-μsec time slot for each PN pulse. Since the 63-chip pulse lasts $0.4 \times 63 = 25.2$ μsec, the PPM is capable of a maximum deviation of ± 49.9 μsec (the deviation derived from each audio sample is applied to the entire 63-chip sequence).

As stated in the Introduction, the PN pulse balance-modulates a 280 Mc/sec carrier signal. Upon reception, the 280 Mc/sec signal is heterodyned to an if of 31 Mc/sec. Quadrature mixers are used to bring the signal down to base-band. The signals at points A and B in Fig. 1 are the base-band binary signals. After hard limiting and sampling, these binary signals are read into their respective digital matched filters. When the sequences are completely read into the DMF's, output pulses occur. These are squared (in the usual mathematical sense) and added to yield a composite spike. This spike is detected by a threshold circuit and is fed into the pulse-position demodulator which is wired, in turn, to the handset.

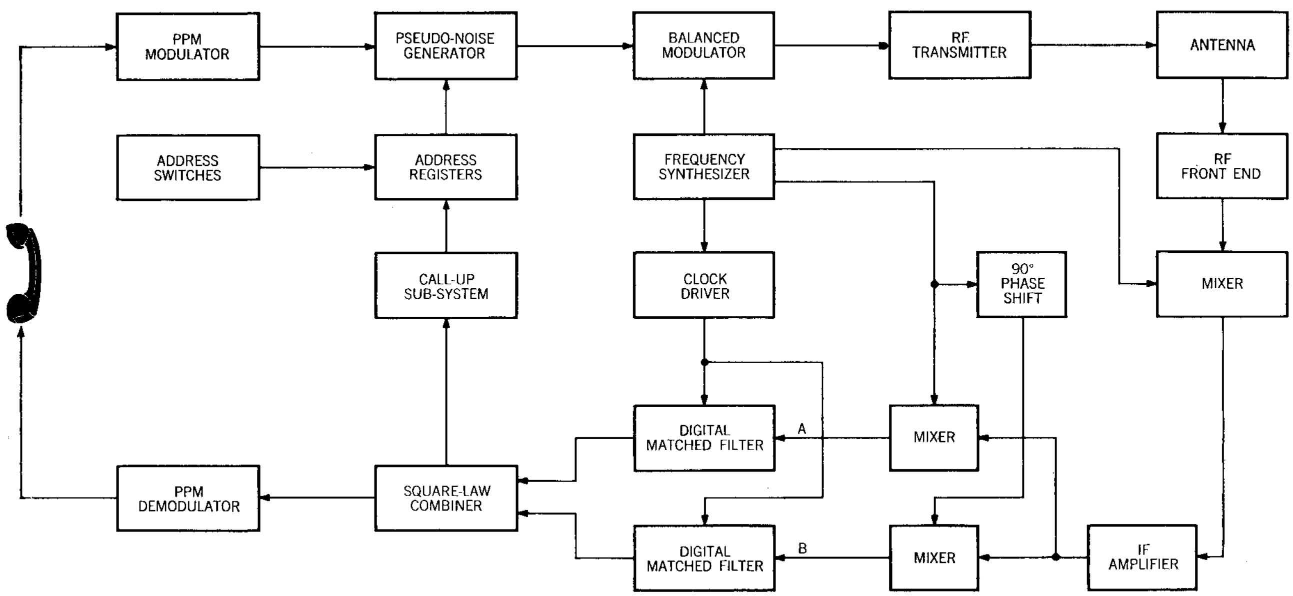
For purposes of discussion, we can divide the radio set into the call-up system, the rf system, and the digital matched filter system. In the following sections, the characteristics of the rather conventional rf system are given, the logic of the call-up system is described, and the operation of the digital matched filters is discussed.

Radio-frequency system

The rf system consists of a balanced modulator, rf transmitter, rf front end, mixer, if amplifier, synchronous demodulator, frequency synthesizer, switchable low-pass filters, and a unipole antenna.

The balanced modulator utilizes the 63-chip pseudo-noise sequences from the PNG to bi-phase modulate a 280 Mc/sec carrier. Utilization of high-speed transistors and ferrite core transformers in the balanced modulator has resulted in simplified construction while yielding good spectral characteristics. The carrier has been suppressed by approximately 23 dB with well-balanced sidebands.

Figure 1 Functional block diagram of pulsed pseudo-noise radio set.



The output signal level from the balanced modulator is -4 dBm.

The transmitter amplifies the PN signal from the balanced modulator to approximately one watt. Double-tuned interstage circuits have been employed, enabling nearly linear phase characteristics over the band of interest.

The rf front end consists of two stages of rf amplification and one mixer stage. The noise figure of the receiver is approximately 4 dB. Double-tuned circuits with neutralized stages provide 13 dB gain per stage at 280 Mc/sec with a bandwidth of approximately 5 Mc/sec. An image trap is included which provides over 40 dB image rejection.

The if amplifier is centered at 31 Mc/sec and provides the additional gain necessary for operation of the synchronous (quadrature) demodulator. A manual gain control is provided to adjust for various operational environments and provides approximately 50 dB control of the if gain. The if amplifier is followed by a hard limiter which enables the system to perform better in the presence of strong interfering co-channel signals. Inasmuch as the system does not operate with an AGC loop, very strong signals may cause the latter if stages to become saturated. This results in some phase distortion if the desired signal is coincident with the strong saturating signal. However, tests performed to date in the laboratory have shown that very strong saturating signals do not cause an audible interference until the input signal level reaches approximately -30 dBm (or approximately 100 dB above the receiver noise power).

The quadrature demodulator consists of two synchronous detectors and a $\pi/2$ phase shifter. The demodulator is operated such that the local oscillator signal at one detector is 90° out of phase with the local oscillator at the other detector. This operation ensures that asynchronous operation may be carried on successfully regardless of received signal phase variations. The operation of each synchronous detector is similar to that of the balanced modulator.

The detected base-band signal from the quadrature demodulator is fed to two bridge switches. Each bridge switch steers the received signal to a low-pass Bessel filter (Bessel filters are employed to provide linear phase characteristics with adequate local oscillator rejection). The cut-off frequencies of the two filters are 100 Kc/sec, which is used during the calling mode, and 2.5 Mc/sec, which is used during the talking mode.

The radio set includes a frequency synthesizer where the various CW signals required for system operation are generated. These signals include a 2.5 Mc/sec square wave used as the system clock, a 31 Mc/sec signal used as the local oscillator in the quadrature demodulators, and the 280 Mc/sec carrier signal. The basic oscillator is crystal controlled and operates at 2.5 Mc/sec.

The antenna employed in the portable VHF system is a unipole device which provides omni-azimuthal coverage. The voltage standing wave ratio (VSWR) of the antenna is approximately 1.4 with an input impedance of 50 ohms. (The loss associated with a VSWR of 1.4 is approximately 0.1 dB.)

Call-up system

Each user is assigned a distinct six (octal) digit address. (The address corresponds to the telephone number employed in standard telephone communications.) The address is inserted into the system via six octal switches on the front panel of the set. The octal number is converted to a binary pattern 17 bits in length. (The 18th available bit is not used.) This binary address is fed into a 17-stage PNG as the initial pattern for the generator. The actual 63-bit PN sequence that is transmitted is determined by the initial 17-bit address that is inserted into the PNG. Therefore, each user is assigned a unique six digit octal number as his own address, and may call any other user by dialing the six digit octal address of the party he wishes to call.

The PNG consists of a 17-stage shift register with several stages tapped and combined in a modulo-two adder (EXCLUSIVE-OR circuit). The output of the adder is fed back to the shift register so that the complete PNG will generate a repetitive (or cyclic) digital sequence. The feedback connections are selected so that the period of the cyclic sequence is the maximum possible consistent with the number of shift register stages employed in the PNG. There are several feedback connections that will produce a maximum length sequence (M-sequence) and these are determined by primitive polynomials. The PNG utilizes two sets of feedback patterns that are automatically selected as a function of the system mode. These patterns are given by the primitive polynomial $x^{17} + x^3 + 1$ and its complement $x^{17} + x^{14} + 1$.

The period of a 17-stage M-sequence is $2^{17} - 1$ or 131,071 bits. The radio set utilizes 63-bit segments of this M-sequence for communications. Therefore, the M-sequence can be divided into $131{,}071/63 \cong 2080$ non-overlapping segments. Inasmuch as the radio set utilizes 63-bit segments of an M-sequence, the autocorrelation function of the sidelobes is not zero. However, an IBM 7094 computer was programmed to compute the auto- and cross-correlations between these segments and showed that a maximum sidelobe of 25% can be obtained with over 50% of these sequences. Therefore, the total number of "good" sequences reduces the number of subscribers to approximately 1,000. A larger number could be attained with a longer shift register. The cross-correlations were computed with both full and partial overlap of the sequences.

The call-up system performs the function of automatically establishing a connection between two users. Since the subscriber who initiates the call knows the address of the user he wishes to call, his system is automatically ready for talking. However, the subscriber who is being called does not know the address of the party who is calling him. Without this information he cannot reach the calling party. It is necessary for the calling party to send his own address to the called subscriber, so that the called party may store this address in his pseudo-noise generator. This is accomplished by the calling party transmitting a digital message conveying his 17-bit address to the called party. Each "1" in the digital message is transmitted using the PNG feedback pattern corresponding to the primitive polynomial $x^{17} + x^{3} + 1$. Each "0" in the digital message is transmitted using the PNG feedback pattern corresponding to the primitive polynomial $x^{17} + x^{14} + 1$. (The feedback loops in the PNG are switched to preclude using two PNG's for this function.) The digital matched filter in the called party's receiver therefore responds to this 17-bit binary message and stores it in his address register for communications during the remainder of the call. After cycling through 63 bits for each pulse, the PNG is reset by the controls of this address register. The data rate and PNG clock rate during the calling mode are decreased to 1 Kc/sec and 100 Kc/sec, respectively, in order to increase the ratio of the pulse energy to the one-sided noise power spectral density at the filter input, and thereby reduce the bit error probability, as is shown during the discussion of the digital matched filter.

The digital matched filter

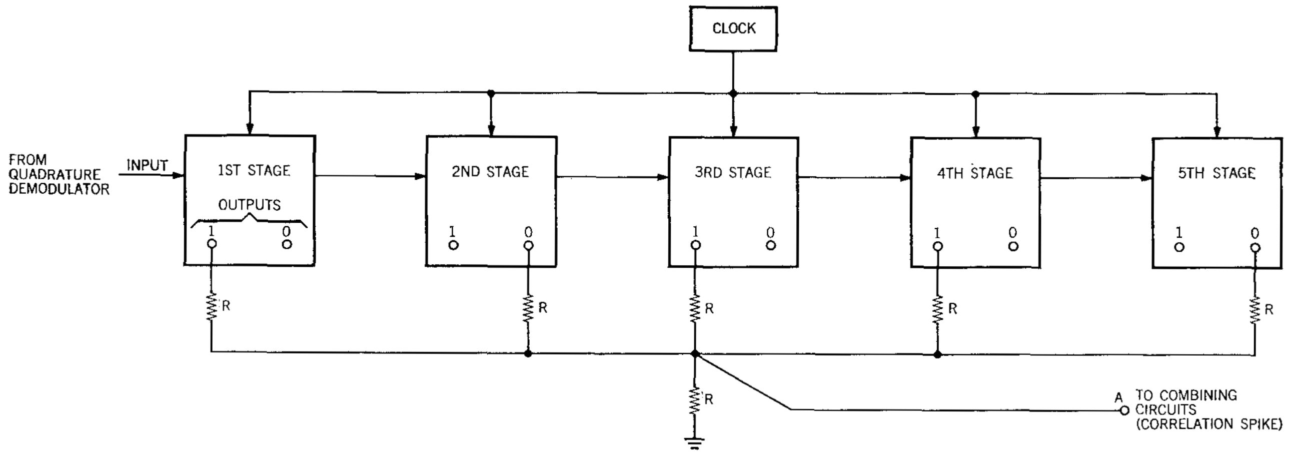
The digital matched filter is the heart of the radio set. This is the device which recognizes that the receiver is being called. It consists of a 63-chip shift register, a clock driver, and a resistive-network adder. For illustrative purposes, Fig. 2 demonstrates how similar elements would be linked together in a 5-chip matched filter. In this illustration the filter's wiring gives it a 5-chip address of 10110.

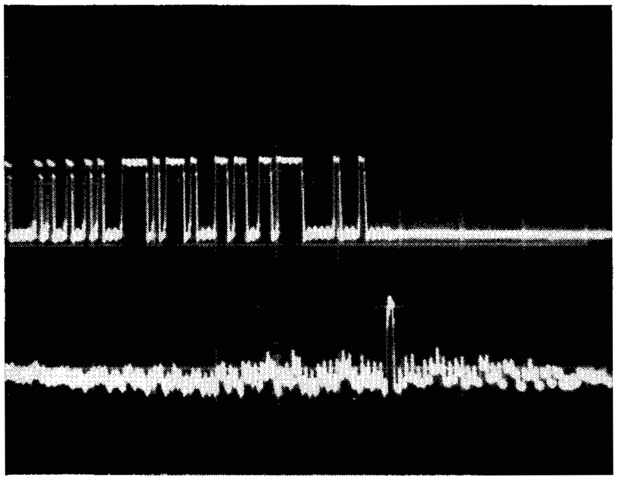
When a trigger (i.e., a shift register stage) contains a "1," the "1" side is at 0 volts and the "0" side at -6 volts. Assuming the trigger voltage to be derived from a source of zero internal impedance, it is easy to show that when the pattern in the 5-stage register is 01001, the voltage at point A is -5 volts. When the pattern is 10110, the voltage at point A is 0 volts. For any other binary pattern the output will be between 0 and -5 volts. In the case of the 63-chip register, the output peak at the instant of match is 0 volts and $-6 \times 63/64 = -5.9$ volts for a perfect mismatch. (This condition occurs when the rf phase difference between the transmitter and receiver exceeds $\pm 90°$.) For a signal which matches the desired signal in about one-half its digits, i.e., nearly zero correlation, the output is approximately -3 volts. The DMF output, therefore, is a bipolar signal whose polarity depends upon the phase of the received rf signal.

The binary pattern, or address, to which the DMF is matched is simply a function of which sides of the triggers are tied to the resistive-network adder. In the present sets, the address can be altered by replacing three printed circuit cards. It would be a straightforward task to make this change by electronic gating.

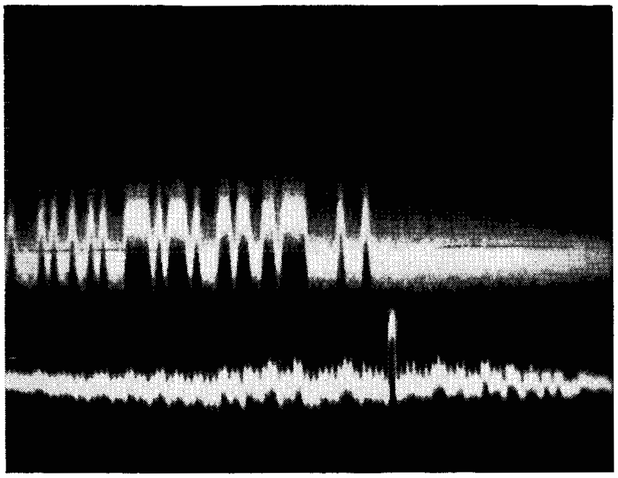
The DMF has been tested with various values of resistors in the adder. Any value above 2 K ohms yields a good response. With sufficiently large resistors, many adders can be driven without the aid of emitter followers. This means that with very little additional expense a single set can respond to many addresses.

Figure 2 A five-stage digital matched filter. The choice of trigger output side for connection to the summing resistor R determines receiver address. The DMF of the radio described in this paper has 63 stages, but is otherwise the same.





(a)



(b)

Figure 3 Oscilloscope traces of DMF response to matching input signal. (a) Noise absent, (b) Noise present.

Figure 3 contains an oscillographic picture of the output of the DMF. The output voltage stays within a few tenths of a volt of -3 volts until the desired binary sequence is in the register. The output then peaks up to 0 volts. Should the sequence be received with a few chips in error, the peak should still be high enough to cross the threshold detector.

It is desirable to operate the radio in a phase-incoherent mode to avoid delays for establishing phase coherence and possible loss of coherence. This is done by using quadrature detection. This kind of receiver has been explained in clear fashion by Elspas.[7] Quadrature detection requires the use of two mixers (see Fig. 1) with local oscillators 90° out of phase with one another. The outputs of these mixers are baseband signals with a bandwidth of about 1.25 Mc/sec, the approximate bandwidth of the base-band pseudo-noise.

Each mixer output is fed to a separate matched filter. The PN signals are sampled and then shifted down the registers. The outputs of the two DMF's are squared and added by the quadrature combining circuit in Fig. 4. This corresponds to recombination of the sine and cosine components of the signal derived from the quadrature detector (See Van Blerkom, et al.[6]). The square law performance is attained by using the current-voltage characteristic of forward biased diodes.

Monostable multivibrators are used as output circuits to serve as threshold pulse detectors and to increase the energy of the detected pulses to drive subsequent circuits. Separate outputs are used for call-up detection and conversation detection, so that these thresholds can be individually adjusted.

If a single DMF is used, the quality of reception alternates between good and bad at a rate determined by the beat frequency between the local if oscillator and that at the remote transmitter. When both DMF's are used with quadrature combining, the effect theoretically disappears. In practice, the local oscillator signals to the mixers must be carefully phased to be 90° apart and the separate channels must be symmetric. Small differences can yield an audible beat.

It should be carefully noted that all of the signal processing prior to the DMF is completely linear. This is important because nonlinearities normally suppress the weaker of two signals. If jamming is a threat, square law detection should be performed only after the processing gain of the DMF has pulled the desired signal level above that of the jammer. Unfortunately, since the shift register is a binary device, it is necessary to hard limit the signals at the input to the DMF. It will be shown below that the loss is at most 2 dB when compared to an ideal linear matched filter.

It was found that using a single clock driver circuit with a high-current, high-speed complementary emitter follower was preferable to using a driving tree for the shift pulses. The tree requires several extra circuits and timing adjustments. For longer shift registers, a tree would have to be used. It is felt that the high-current circuit should drive at least 100 stages without resorting to a tree.

A sampling problem arises in the design of a DMF. Since the binary signals are band-limited, their rise and fall times are finite. When the 2.5 Mc/sec sampling clock at the receiver is not synchronized with the one at the remote transmitter, the DMF will sometimes be sampling near the zero-crossing of the binary data. At such times the output spike of the DMF will be more easily reduced by noise, and cancellation errors will become more likely.

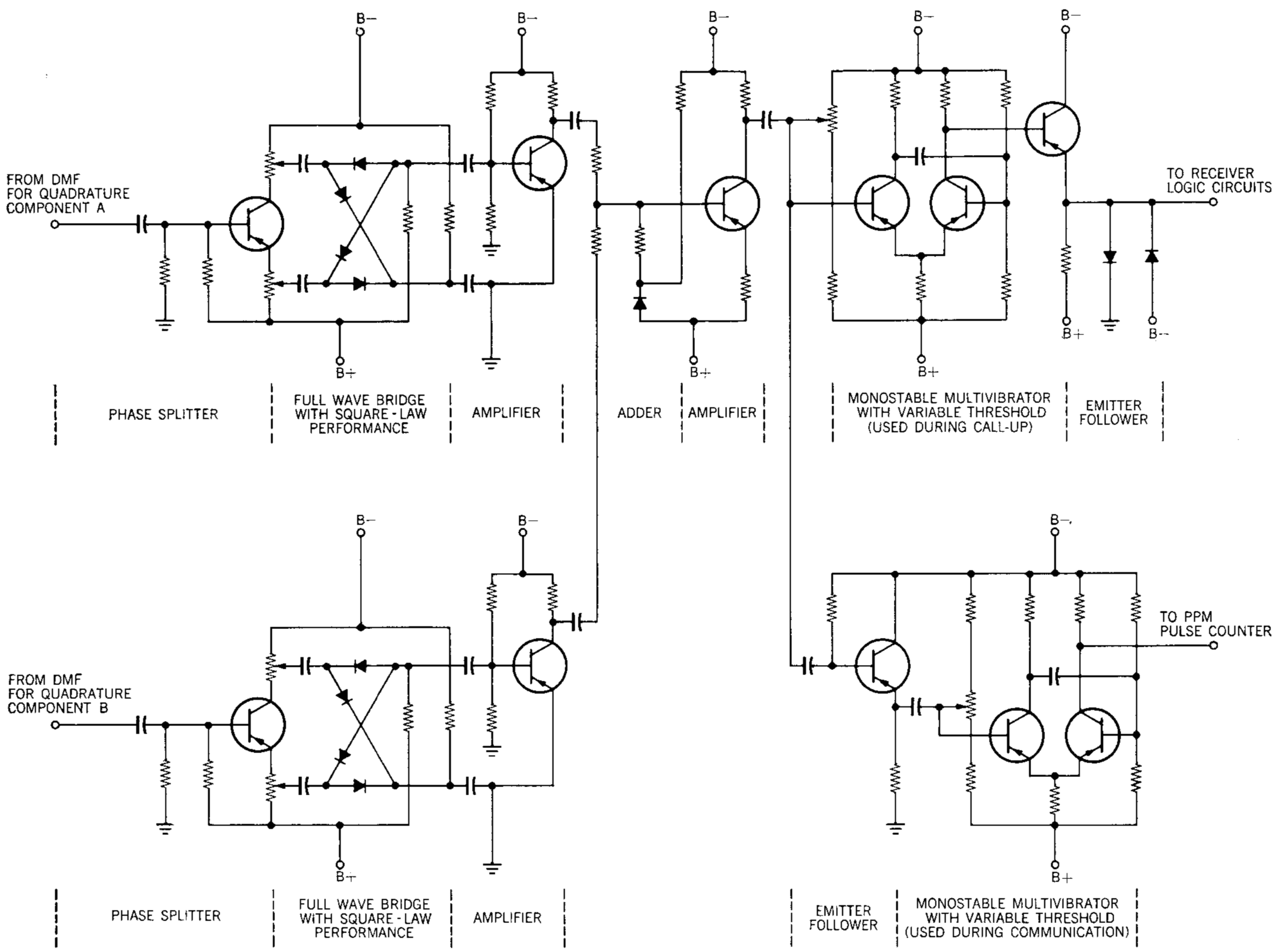
Effectively, the signal-to-noise ratio at the input of the DMF will vary as the beat frequency between the local 2.5 Mc/sec shifting clock and that at the remote transmitter.

This problem causes an annoying acoustical disturbance. It is due simply to the sampled nature of the signal in the DMF. Several different solutions have been tried including frequency offset of transmitter and receiver clocks, sample jitter and simple clock sync mechanisms. The technique selected for implementation with the system is a delay-line integrator. The delay time is equal to the width of the correlation spike from the DMF (0.4 μsec). This system has effectively eliminated the acoustical disturbance and is presently installed and operating with the system.

Maximum pulse deviation is a very important design parameter. As the thermal noise increases, narrower deviations are required. In order to achieve the advantage of narrow deviation, a gating circuit is required at the DMF output. The circuit does not allow noise pulses to pass during those portions of the 125-μsec interval which cannot be occupied by signal pulses. The gate has been designed to synchronize to the nominal center of a gate interval. It uses only the PPM pulses themselves and no special synchronizing information. Another advantage of this gate is that any strong interfering pulses arriving during the blanked intervals do not interfere with the transmission. The gate generator consists of a sampled-data phase-locked loop and requires very little time for synchronization (nominally less than 10 msec).

Carrier phase synchronism between the local if oscillator and the incoming if carrier signal is not necessary. This is due to the use of quadrature mixers and the DMF combiner circuit. Nevertheless, the frequency offset between local oscillator and signal carrier is limited. For example, if the frequency offset is equal to the reciprocal of the pulse length (i.e., 40 kc/sec), the output of the DMF's in the absence of noise is zero. An offset

Figure 4 Quadrature combining circuit.



of 10 kc/sec or less results in negligible loss. Indeed, since the set functions as a radar set, from the signal processing point of view, the usual ambiguity function plots are applicable.

Finally, it was decided to use a digital shift register instead of an analog delay line for the following reasons:

1. There is no known limitation to the length of the register with digital techniques. In some applications a register of a length (equivalent to the bandwidth-time product of the pulse) much greater than 63 is required.
2. The shifting rate is easily varied.
3. There is no signal attenuation as the sequence is shifted down the register.
4. Integrated digital circuits alleviate greatly size, weight and power problems.
5. The DMF register can be used to perform other functions when not operating as a filter. (For example, one of the DMF's in the system is used to store its own address for transmission to the party being called.)
6. Simple resistor matrix gating arrangements can be used in applications where variable addresses are required.

Comparison of digital and linear matched filter operation

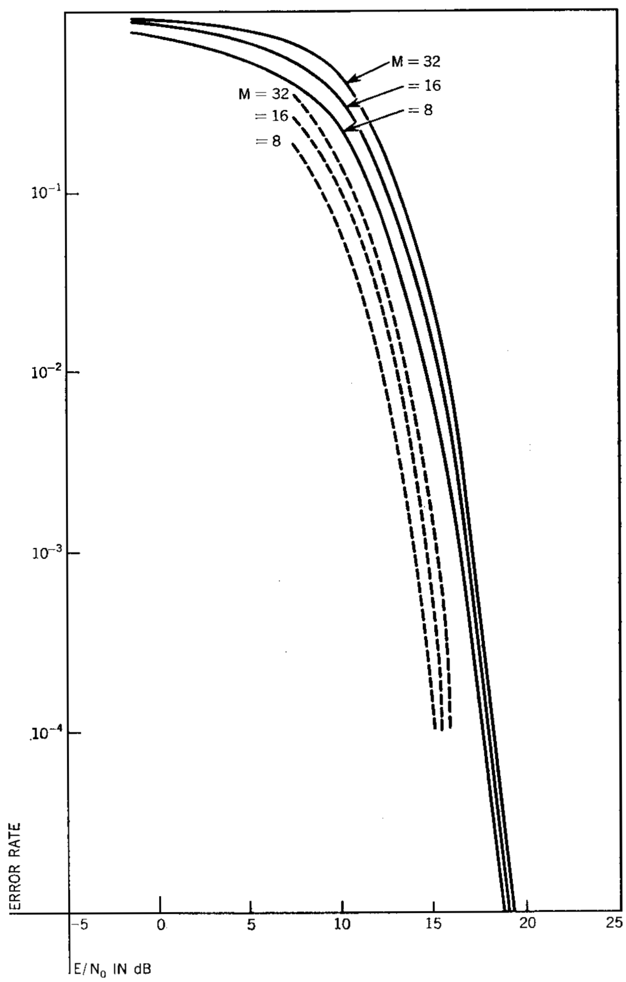
The digital matched filter makes chip-by-chip decisions as the PN sequence is read in to the shift register. This constitutes a non-linear operation. A theoretical comparison of the DMF with a linear matched filter operating in an M-ary system has been made. The calculations involved only the base-band problem. That is, additive white Gaussian noise at the input to the DMF was considered. Error rate vs E/N_0 for a 64-digit sequence is plotted in Fig. 5. Error rate is the probability that the output from the DMF occupies the wrong position, E is the pulse energy of the PN sequence, and N_0 is the one-sided noise power spectral density of the input to the filter. (The ratio E/N_0 is a basic limiting parameter on detectability in pulse communications.) Although the actual receiver described in this paper uses threshold detection for simplicity, the curves of Fig. 5 were calculated assuming "greatest-of" detection. Using circuits developed by Wiggins,[8] one can modify the threshold detector so that it performs nearly as well as a "greatest-of" detector.

The number of positions which the output pulse may assume is finite because a pulse can occur only when the register is shifted at intervals of 0.4 μsec. Effectively, this is a quantized system. For example, if it is desired to have 32 quantizing levels, the gate at the DMF output should be open for $0.4 \times 32 = 12.8$ μsec. The maximum pulse deviation for this operation would be ± 6.4 μsec. The theoretical curves were derived by considering a hypothetical receiver which tests all M possible quantizing levels ($M = 8, 16, 32$ were considered) using "greatest-of" detection.

In addition, some curves were calculated for sequences of length $N = 32$, and 128 chips, with almost negligible differences from the curve for $N = 64$. Independent chip decisions and zero crosscorrelation were assumed. The calculation for the DMF was carried out on an IBM 7094 computer. The curves for the linear filters were derived from well-known relationships for linear matched filters. The curves show very little difference (at most 2 dB) in the performances of the DMF and the linear filter. Tests performed to date indicate results 2 to 3 dB worse than the predicted values.

The quantized PPM is essentially a higher-order (M-ary) alphabet. This is a simple implementation of the M-ary concept whose advantages have been pointed

Figure 5 Error rate vs E/N_0 for theoretical 64-chip matched filters at three quantizing levels, M. Solid lines: non-linear digital matched filter; dashed lines: ideal linear matched filter.



out by Turin[9] and Reiger.[10] The elements of the alphabet consist of shifted versions of the PN sequence. In effect, the crosscorrelation of members of the alphabet is the autocorrelation of the sequence. Since chip-by-chip decisions are made, it is possible to interpret the system as a digital error-correcting machine. Nevertheless, analysis using the M-ary alphabet interpretation is simpler and is valid if one takes account of the usual hard-limiting losses in the presence of white noise and other interference.

Conclusions

A radio set has been described which uses pulse position modulation of pseudo-noise pulses. The implementation has made use of the advantages inherent in higher-order signaling alphabets. Moreover, pulsing of the pseudo-noise signal enables many disturbances to be gated out of the receiver. Intelligible signals can be achieved even when the received rf pulses are essentially masked by the noise. In particular, the system has provided intelligible audio even with an input signal-to-noise ratio of -8 dB.

It is possible to operate many sets in the same frequency channel because each set responds only to a signal containing its unique address. Analysis and simulation have shown that the system can support 13 equal-power talkers per megacycle at an audio signal-to-noise ratio of 13 dB. Analysis also shows that the theoretical performance of a digital matched filter is comparable to that of a linear filter, thus eliminating the disadvantages of a tapped delay line.

The use of a maximal length sequence generator has made it possible to generate a 63-digit address with only 6 octal digits. A simple method for sending address data during call-up has been achieved by using complementary polynomials.

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