# FDM-FM COCHANNEL INTERFERENCE IDENTIFICATION SYSTEM

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## ABSTRACT

A prototype interference identification system has been developed to detect and identify interfering FDM-FM carriers originating within the INTELSAT* system. Interfering carriers are identified by distinctive code (signature) modulation of the energy-dispersal waveform of each FDM-FM carrier. Identification presently is accomplished within 10 minutes for ratios of interfering carrier power to noise power down to -2 dB, and for ratios of interfering carrier power to desired carrier power down to -17 dB. Possible improvements are discussed for more rapid identification.

## I. INTRODUCTION

In the era of INTELSAT V satellites, extensive frequency reuse is expected because of growing space and polarization diversities and because of the growing number of satellite and terrestrial microwave communication systems. Frequency reuse increases the probability of cochannel interference. Since the interferer generally is unaware of his effect on other signals, and since FM carrier cross-talk generally is unintelligible (and hence not

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* International Telecommunications Satellite Organization

readily traceable), it is becoming increasingly critical to be able to detect FDM-FM interference and to identify sources of interference.

This paper presents an FDM-FM interference identification technique that has been developed, implemented, and successfully tested. The basic concept of the developed FDM-FM interference identification technique is to provide each FDM-FM carrier with a unique signature through distinctive encoding of each individual carrier's energy dispersal waveform (EDW) at the FM transmitter. At the FM receiver, the EDW signatures (or codes) of the desired carrier and all in-band interfering carriers remain unambiguous and can be recovered through bandpass filtering and envelope-detecting the FM demodulator output. By cross-correlating the recovered codes with locally generated replicas of all candidate codes, interference identification is accomplished when high cross-correlation is found between a recovered code and a particular locally generated code. This cross-correlation technique is essentially the same as the standard delay-locked loop (DLL) technique used in spread spectrum communications receivers. [1,2,3,4]

The two most critical aspects of this developed technique are encoding of the EDW at the FM transmitter and detection of the EDW codes at the FM receiver. Sections II and III address these two topics. Section IV describes the implemented test system. Section V summarizes test results, and Section VI presents conclusions and discussions. Detailed performance analysis is provided in the appendix.

## II. ENCODING OF THE ENERGY-DISPERSAL WAVEFORM

Before considering encoding of the energy-dispersal waveform (EDW) in detail, we review briefly the function of the conventional EDW.

In the INTELSAT FM telephony carriers, many baseband voice signals (a few to over a thousand) are frequency-division-multiplexed (FDM) together at a frequency above 12 kHz, as shown in Figure 1. The resulting baseband signal then frequency-modulates the carrier. Consequently, the modulating signal level and the carrier RF bandwidth vary continuously with the number of active voice circuits. During periods of low usage (e.g., at night), the carrier RF bandwidth is small. Since the carrier RF power is relatively invariant, the resultant power spectral density is high. This high power spectral density tends to interfere with other communciations. To reduce such interference, a conventional EDW, which is an adjustable-level low-frequency (20 to 150 Hz) triangular wave, is added to the unused frequency region (0 to 12 kHz) of the baseband signal to sweep the FM carrier center-frequency (and thus to disperse the carrier power spectral density) across the full allocated RF bandwidth of the carrier. As the voice traffic changes, the EDW amplitude must be adjusted accordingly in order to maintain constant RF frequency occupancy and hence constant power spectral density.

With this background in conventional EDW, we now consider the encoding of each FDM-FM carrier's EDW to provide an identification signature. Gold codes,[5,6,7] having properties similar to the pseudo-random noise (PRN) codes, were selected for EDW encoding because of four reasons. First, each family of Gold codes has a large number (e.g., 1023 or more) of distinct codes available for unique code assignment to all carriers in the system. Second, Gold codes are periodic and can be generated easily by two shift registers with feedback taps. Third, a large code distance (or low cross-correlation) exists between different Gold codes, thereby providing high protection against false code identification (hence against false interference identification). Fourth, because of Gold code similarity to the PRN code, the widely used delay-locked loop[1,2,3,4] can be employed (a) to perform the necessary cross-correlation function, (b) to synchronize the local reference Gold code with the recovered Gold code at the receiver, and (c) to facilitate the interference identification unit design.

The simplest EDW encoding convention is to make the Gold code bit interval equal to half of the EDW triangular wave period, with the code transitions synchronized to the triangular wave zero-crossings. Each Gold code "one" bit becomes a positive triangle, and each Gold code "zero" bit appears as a negative triangle in the coded EDW. Figure 2 shows the uncoded EDW, a segment of Gold code, and the coded EDW. The functional block diagram of the EDW encoder and an implemented EDW encoder are shown in Figures 3 and 4, respectively.

## III. RECOVERY AND DETECTION OF EDW SIGNATURE

Unlike the linear amplitude demodulation case, two interfering FM carriers do not yield the baseband signals of both carriers at the frequency demodulator output.[8] In the typical weak interference situations as shown in Figure 5, the FM demodulator output consists of the desired carrier's basband signal plus a series of convolutional interference components, one for each weak FM interferer.[9,10] Each convolutional interference component, centering at the difference frequency of the desired and interfering carriers, has a baseband power spectrum given by the convolution of the desired and interfering RF power spectra, and has a power level proportional to the interfering-to-desired carrier power ratio. More importantly, each convolutional interference component is envelope-modulated by the interferer's coded EDW as illustrated in Figure 5. By passing this wideband convolutional interference component through a narrow bandpass filter and then an envelope detector, the interferer's coded EDW can be recovered for signature identification.

For high signature identification sensitivity, the recovered interferers' EDW codes are cross-correlated with local replicas of all candidate codes. Interference detection is indicated when high correlation exists between the recovered code and a particular local candidate code. To synchronize and track the local code epoch with the received code epoch and to perform the cross-correlation detection, a conventional delay-locked loop (DLL) is employed. The detection performance analysis is detailed in the appendix.

## IV. LABORATORY TEST

Two STI 2002 coded EDW generators are used at the FM transmitters to generate coded EDW's for the desired and interfering carriers. The family of Gold codes having a 1023-bit period and selective bit rate between 40 and 1200 bps was implemented in the STI 2002 generator.

The laboratory signature-recovery system is shown in Figure 6. The frequency-demodualtor was a standard multi-channel carrier demodulator module. The predetection bandpass filters and the envelope-detector were a spectrum analyzer operating in the zero-span (non-sweeping) mode. This provides frequency-tunable bandpass filtering, envelope detection, video filtering, and visual monitoring, all in one package convenient for test. In the field installation, however, less expensive dedicated equipment could be used for the functions presently performed by the spectrum analyzer.

The interfering carrier's EDW signature is identified by the STI 2001 interference identification unit shown in Figure 7. This is a microprocessor-controlled digital implementation of a delay-lock loop code correlator, as used in most direct sequence spread spectrum communciation and navigation receivers. One unit was designed and built as a prototype for laboratory and field use. A local Gold code generator in the interference identification unit, generates a reference code identical to the signature code of one of the suspected interfering carriers. Under microprocessor control, the reference code is time slewed slowly relative to the received signature code to check for synchronization and correlation. If high cross-correlation is detected (i.e., if the codes appear to be similar), then the reference code is delay-locked to the received code (i.e., its epoch delay is locked to the value which yields maximum cross-correlation). If the received and reference codes continue to show high cross-correlation for a predetermined (adjustable) time, then the reference code sequence is considered to be identical to the received code sequence; i.e., the received carrier is identified. The designation (name) of the interfering carrier is determined simply by looking up the code assignment table to see which transmit carrier is assigned to use the received code. If, as often occurs, continued cross-correlation monitoring shows that the high correlation was only temporary, then, still under microprocessor control, delay-lock is broken, and the reference code again slews with respect to the received code. If identification is not made within the time sufficient to slew the reference code epoch past entire period of the received code, then it is concluded that the reference code sequence is different from that of either of the received carriers; in that case, the reference code generator is set to a different address (code sequence), and the correlation measurement is repeated.

In an interference situation, at least two signature codes (one from the desired carrier, and one from the interfering carrier) are superposed in the envelope detector output. In the signature recovery system described above, the amplitudes of the two signature codes (at the correlation detector input port) are relatively independent of the levels of the two RF carriers. However, the signature code amplitudes do depend on EDW magnitudes or the amounts of energy-dispersal used on the two carriers. Thus for signature identification, the worst situation is a slightly-dispersed carrier interfering with a highly-dispersed desired carrier. In this situation, the correlator must be able to identify the interference carrier's weak EDW signature in the presence of the desired carrier's strong EDW signature, implying the requirement for low cross-correlation between distinct EDW codes. Recall from Section II that this was one of the reasons that Gold codes were selected for EDW encoding.

False lock is minimized by making the reference code slew rate small, and integrating the cross-correlation measurement over a large number of code bits (up to the full 1023-bit period of the code sequence). This tends to require a long time (about 10 minutes) for code cross-correlation, hence for interference identification. Several methods have been tested for reducing the identification times, as described below.

First, the correlation threshold must be set high enough to minimize false lock, but not so high as to miss true lock.

Second, the signature code bit rate (hence the energy-dispersal frequency) should be as high as possible in order to allow integration over a large number of code bits (in cross-correlation measurement) during a short time. If the energy-dispersal frequency is excessive, energy-dispersal signal harmonics could interfere with voice or data signals in the baseband shown in Figure 1. It is believed, however, that the code chip rate easily can be 1,500 bits/second, rather than the 40 to 300 b/s that would be required to encode the present INTELSAT energy-dispersal triangular waves of 20 to 150 Hz. In test, code bit rates up to 1,200 b/s were used successfully.

Third, all code generators world-wide should be synchronized in order to avoid having to slew the reference code past the entire 1023-bit code period of the received code in searching for correlation. With present timing system technology, it is simple to maintain synchronization within 0.1 second by resynchronizing only weekly. Such synchronization was used successfully in the laboratory to minimize test time, and the reference code generator (in the interference identification unit) is built with variable code epoch delay to control the search range during cross-correlation.

During test, two adjustments were found to be critical to detection time performance. First, as described above, the correlation threshold must be low enough to identify a weak EDW signature of the interfering carrier in the presence of a strong EDW signature of the desired carrier, yet high enough to avoid excessive temporary false lock. Second, the bandpass filter center-frequency tuning and filter bandwidth setting are critical. Algorithms were found during test for making these adjustments properly.

## V. TEST **RESULTS**

The laboratory system was tested over a wide range of combinations of interference-to-desired carrier power ratio, carrier-to-noise density ratio, EDW magnitude in each carrier, and carrier center-frequency separation. Figure 8 illustrates a recovered EDW at the envelope detector output. Figure 9 shows the typical signature-detection limit as a function of desired carrier-to-interference carrier power ratio and desired carrier-to-noise power ratio. Signature detection was accomplished for combinations of these two parameters in the lower right quadrant of the plot, and not accomplished in the other quadrants. Detection was found to be possible under all typical interference conditions.

## VI. CONCLUSIONS AND DISCUSSIONS

By assigning to each FDM-FM transmit carrier a distinct EDW code, and detecting the recovered EDW codes at the receiver, identification of cochannel FDM-FM interference has been accomplished. According to test results obtained from a developed prototype system, identification within 10 minutes is possible for interfereing carrier-to-noise power ratio as low as -2 dB and for interfering carrier-to-desired carrier power ratio as low as -17 dB.

Three methods to reduce the current 10-minute detection time were described in Section IV. A fourth method, which was not tested during this study, is discussed here. The method consists of recording one full code sequence in real time (about 6 seconds), and then performing the cross-correlation off-line at a much higher rate, hence in a much shorter time. In the usual (real-time) method of on-line correlation, each arriving code bit is used only once, and then is discarded. When the received code bit rate is low, as in the developed interference identification system, this is extremely wasteful of time. It is estimated that cross-correlation time can be shortened by about two orders of magnitude by sampling the correlator input signal, quantizing the samples and storing in random-access memory, and then reading and correlating at high rate. A similar technique has been used successfully for many years in spectrum analyzers for low-frequency signals.

## VII. ACKNOWLEDGEMENTS

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# APPENDIX

# PERFORMANCE ANALYSIS

## FM DEMODULATOR OUTPUT

In the simplified receiver block diagram in Figure 1-A, let $f_o$ and $f_i$ denote the center frequencies of the desired and the ith interfering carriers. The existence of K interfers located within the allocated bandwidth $B_{ao}$ of the desired carrier implies

$$\left| f_i - f_o \right| < B_{ao}/2 \tag{1-A}$$

for i = 1 to K. Assume that the predemodulation IF filter centering at $f_o$ has a bandwidth $B_{IF}$ just wide enough to allow passage of the desired carrier without significant distortion, namely,

$$B_{IF} = B_0 \tag{2-A}$$

where $B_o$ is the occupied bandwidth of the desired carrier. Under small in-band interference ($C >> I_i$) and weak in-band noise conditions which usually apply in our cases of concern, the frequency-demodulator output z(t) is equal approximately to

$$z(t) \simeq k_d \left[k_m x_o(t) + k_m e_o(t)\right] + \sum_{i=1}^{K} y_i(t) + n_z(t) \tag{3-A}$$

where

$k_d$ = FM demodulator constant (volts/Hz)

$k_m$ = FM deviation constant (Hz/volt)

$x_o(t)$ = baseband modulation of the desired carrier of bandwidth $B_{bo}$

$e_o(t)$ = coded EDW or of the desired carrier (4-A)

$$y_i(t) = k_d \sqrt{I_i/C}\; g_i(t) \cos\left[2\pi \left([f_i - f_o]t + f[x_i(t) - x_o(t) + e_i(t) - e\,(t)]dt\right.\right. \tag{5-A}$$

$$g_i(t) = k_d \sqrt{I_i/C}\left(f_i - f_o + k_m \left[x_i(t) - x_o(t) + e_i(t) - e_o(t)\right]\right) \tag{6-A}$$

$x_i(t)$ = baseband modulation of the ith interferer

$e_i(t)$ = coded EDW or IIW of the ith interferer of bandwidth $W_{ei}$

$n_z(t)$ = additive noise with power spectral density $S_{nz}$ (f) at the FM demodulator output

and

$$S_{nz}(f) = \begin{cases} (N/C)\ (k_d\ f)^2/(2B_o)\ , & f < B_{IF}/2 \\ 0\ , & f > B_{IF}/2 \end{cases} \tag{7A}$$

C/N = desired carrier-to-noise power ratio over the occupied bandwidth of the desired carrier.

The noise power spectrum at the output of the FM discriminator is shown in Figure 2-A.

The power spectral densities of various terms in the FM demodulator output z(t) are illustrated in Figure 3-A.

**ENVELOPE DETECTOR OUTPUT**

Among the various terms in the FM demodulator output, only $y_i(t)$ contains the signature $e_i(t)$ of the ith interferer. In general, $y_i(t)$ is a wideband signal because it is frequency-modulated by the modulating signals of the desired and the ith interfering carriers. Expansion of $y_i(t)$ shows that $e_i(t)$ amplitude-modulates the cos $2\pi F_i(t)t$ carrier. Recovery of $e_i(t)$ therefore can be achieved by passing z(t) through a narrowband predetection bandpass filter (BPF) followed by an envelope detector as shown in Figure 4.

Let $\nu_i$ and $W_i$ denote the frequency and bandwidth of the ith predetection BPF. To allow passage of $e_i(t)$ without significant distortion while minimizing the noise effect, $W_i = 2W_{ei}$ should e selected. Under this condition, the BPF acts essentially as an IF to baseband frequency translator. When Fi(t) falls within the BPF bandwidth, i.e., $\nu_i - W_i/2 \le F_i(t) \le \nu_i + W_i/2$, we get an undistorted envelope, $e_i(t)$ at the envelope detector output, if $\nu_i >> W_i$. The requirement for this condition can be seen by noting that

$$k_d\ \upsilon_i \cos 2\pi\upsilon_i t = k_d\ (\upsilon_i - k_m e_i)\ \left(1 + \frac{k_m e_i}{\upsilon_i - k_m e_i} \cos 2\pi\upsilon_i t\right)$$

A well defined envelope can be detected when

$$\frac{k_m e_i}{\nu_i - k_m e_i} \ll 1 \ ,$$

or $\nu_i \gg k_m e_i$; then the above condition is satisfied, and the detected envelope has value of approximately $k_k d_m e_i$.*

The fraction of time that $g_i(t)$ falls within the perdetection BPF bandwidth is given by $W_i/(B_i + B_o)$ where $B_i + B_o$ is the total bandwidth of the convolutional noise, and $B_i$ and $B_o$ are the RF bandwidths of the interference and desired carriers. Noting that $y_i(t)$ passes through the BPF when and only when $F_i(t)$ falls within the BPF bandwidth, we conclude that $y_i$'s BPF output to input power ratio is also equal to $W_i/(B_i + B_o)$. With this, the BPF output can be expressed as

$$h_i(t) = \sqrt{I_i/C}\,[A_{ei}\hat{e}_i(t) + A_{eo}\hat{e}_o(t) + A_{xi}\hat{x}_i(t) + A_{xo}\hat{x}_o(t)] + n_i(t) \qquad \text{(8-A)}$$

where $e_i$, $e_o$, $x_i$, and $x_o$ are normalizing waveforms such that

$$\overline{\hat{e}_i^{\,2}} = \overline{\hat{e}_o^{\,2}} = \overline{\hat{x}_i^{\,2}} = \overline{\hat{x}_o^{\,2}} = 1$$

$$A_{ei}^2 = k_d^2\,(k_m^2\ \overline{e_i^2})\ W_i/(B_i + B_o)$$

$$A_{eo}^2 = k_d^2\,(k_m^2\ \overline{e_o^2})\ W_i/(B_i + B_o)$$

$$A_{xi}^2 = k_d^2\,(k_m^2\ \overline{x_i^2})\ W_i/(B_i + B_o)$$

$$A_{xo}^2 = k_d^2\,(k_m^2\ \overline{x_o^2})\ W_i/(B_i + B_o)$$

The noise $n_i(t)$ is contributed by $x_o(t)$, $n_z(t)$, the undesired components in $g_i(t)$, and $g_i(t)$'s $(j \neq i)$. Inspection of the power spectral densities in Figure 3-A indicates that the effective noise $n_i(t)$ at the envelope detector output depends greatly upon the choice of the center frequency $\nu_i$ of the predetection filter. Increasing tends to decrease the effective noise contributed of $S_{xo}(f)$ but increase the effective noise contributed of $S_{nz}(f)$. To facilitate

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* The value of $\nu_i$ has been typically chosen to be

$$[(f_i - f_o) + \sqrt{3}\,k_m\,\sqrt{\overline{e_i^2}}\,]$$

where a uniform probability distribution function for $g_i$ (t) has been assumed.

computing the effective noise $n_i(t)$ at the envelope detector output, consider the case of the single interferer. Assume that $S_{xi}(f)$ can be approximated by

$$S_{xi}(f) = \frac{k_d^2 k_m^2 \overline{x_i^2}}{B_{bi}} \left(\frac{\sin(\pi f/B_{bi})}{\pi f/B_{bi}}\right)^2 \quad ; \; i = 0, 1, \ldots, K \tag{9-A}$$

where

$k_m^2 \overline{x_i^2}$ = mean squared frequency deviation due to message modulation

$B_{bp}$ = top baseband frequency of the interfering signal

Then it can then be shown that $n_i(t)$ is a baseband, nearly-white noise process with bandwidth = $W_i/2$ and noise power of

$$N_i = W_i n_i/2 \tag{10-A}$$

where the two-sided noise power spectral density $n_i$ is given by

$$n_i(\nu_i)/2 = (k_d k_m)^2 (1 + I_i/C)\, S_{xo}(\nu_i) + S_{nz}(\nu_i) + S_{xi}(\nu_i) \tag{11-A}$$

Note that the total noise power in the predetection BPF is found by integrating the above over the BPF frequency range.

## DETECTABILITY ANALYSIS

As illustrated in Figure 4-A, the envelope detector output $h_i(t)$ is fed into an IIW detector for identifying the interferer's parameters. The IIW detector uses a delay lock loop to correlate a replica of $e_i(t)$ with the envelope detector output.

With a perfectly synchornized local IIW $e_i(t)$, the integrator output $q_i(T)$ becomes

$$q_i(T) = \sqrt{I_i/C}\,(A_{ei} + A_{eo}\, C_e) + n_o(T) \tag{12-A}$$

where

$c_e = \frac{1}{T}\int_0^T \hat{e}_i(t)\, \hat{e}_o(t)\, dt$ = cross-correlation level between IIW's when T equals to the period of the IIW's

$$n_o(T) = \frac{1}{T}\int_0^T n_e(t)\, dt$$ = sampled noise voice voltage at the integrator output

is a random variable, and

$$n_e(t) = \sqrt{\frac{I_i}{C}}\,\{[A_{xi}\; x_i(t) + A_{xo}x_o(t)] + n_i(t)\}e_i(t) \qquad (13\text{-A})$$

is an additive noise of two-sided power spectral density

$$S_{ne}\,(f) = n_i/2\,.\,\mathrm{sinc}^2\,(f/W_i) \qquad (14\text{-A})$$

The variance of $n_o(T)$ can be shown to be given by

$$\sigma^2 = \begin{cases} n_i/(2T) & \text{for } 1/T \ll W_i \\ n_i/(2T^2W_i) \quad (\gg n_i/2T) & \text{for } 1/T \gg W_i \end{cases} \qquad (15\text{-A})$$

where $W_i/2$ and $l/(2T)$ can be regarded as predetection and postdetection bandwidths, respectively. In most practical cases of applciation, $l/T \ll W_i$ such that

$$\sigma^2 = n_i/(2T) \qquad (16\text{-A})$$

Detection of the ith IIW is declared when $q_i(T)$ exceeds a preset threshold $\alpha$. For simplicity, assume that $n_o(T)$ and hence $q_i(T)$ are Gaussian. The probability of detection $P_D$ and false alarm $P_F$ can then be shown to be equal to

$$P_D\,[q_i(t) > \alpha/x_i \text{ and } e_i \text{ present}] = \tfrac{1}{2}\,\mathrm{erfc}\;\frac{\alpha - \sqrt{I_i/C}(A_{ei} - A_{eo}\,C_e)}{\sqrt{2}\,\sigma} \qquad (17\text{-A})$$

$$P_F\,[q_i(t) > \alpha/x_i \text{ and } e_i \text{ absent}] = \tfrac{1}{2}\,\mathrm{erfc}\;\frac{\alpha}{\sqrt{2}\,\sigma} \qquad (18\text{-A})$$

In the actual detector design, the threshold $\alpha$ is set based upon the expected noise standard deviation and cross-correlation level $C_e$ in order to achieve a desired probability of false alarm. The probability of detection is then a function of $I_i/C$, $\alpha$, $C_e$, and $\sigma$.

**LEGEND**

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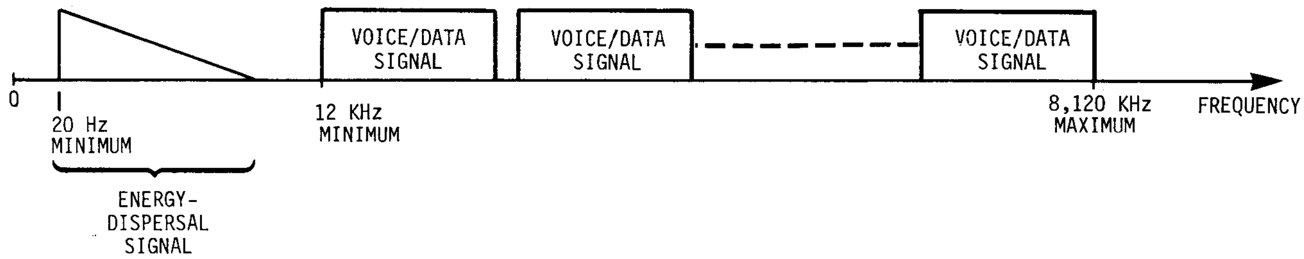



**FIGURE 1 BASEBAND FREQUENCY-DIVISION MULTIPLEXING (FDM)OF VOICE AND DATA SIGNALS FOR A TYPICAL FM CARRIER**

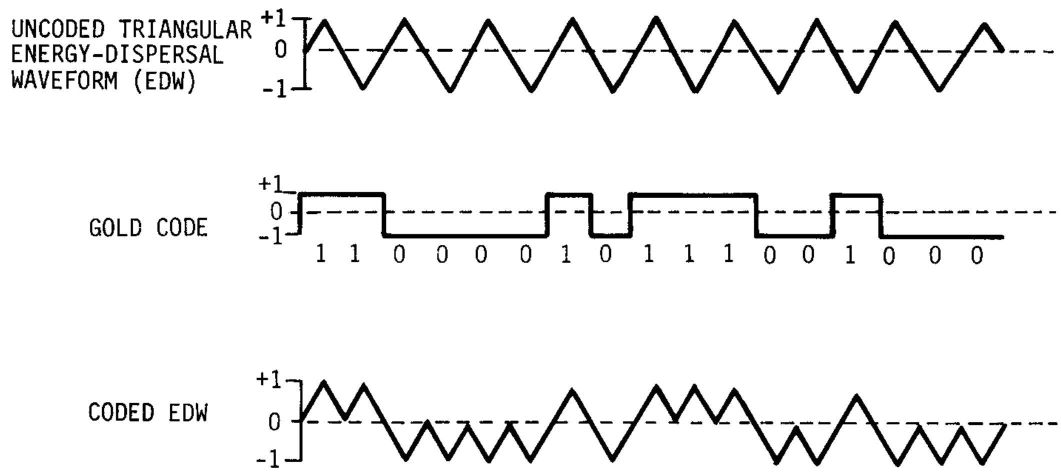



**FIGURE 2 GOLD CODE AND CODED ENERGY-DISPERSAL WAVEFORM (EDW)**

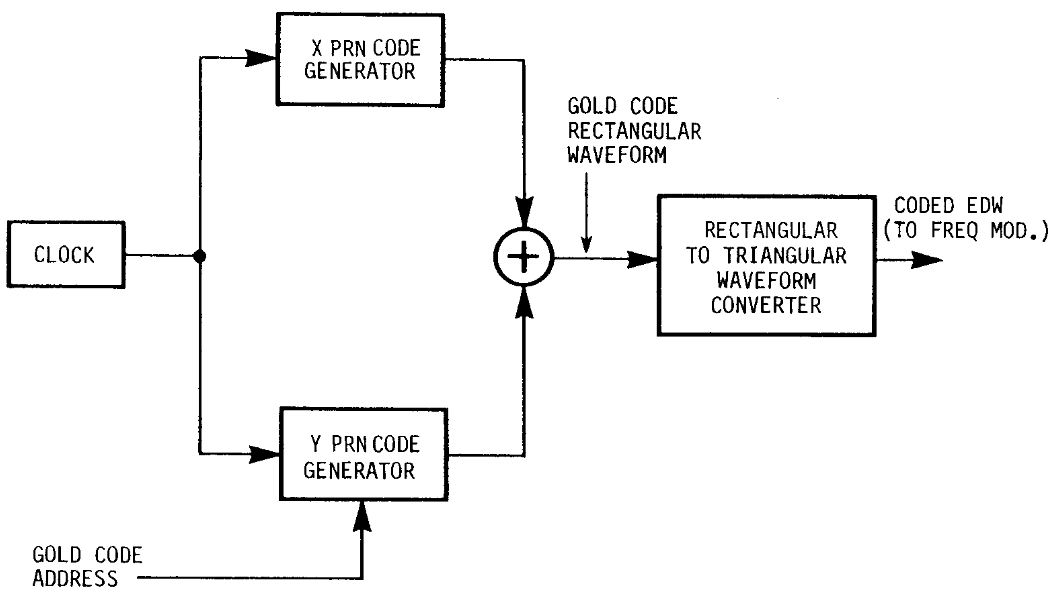



**FIGURE 3 FUNCTIONAL BLOCK DIAGRAM OF CODED EDW GENERATOR**

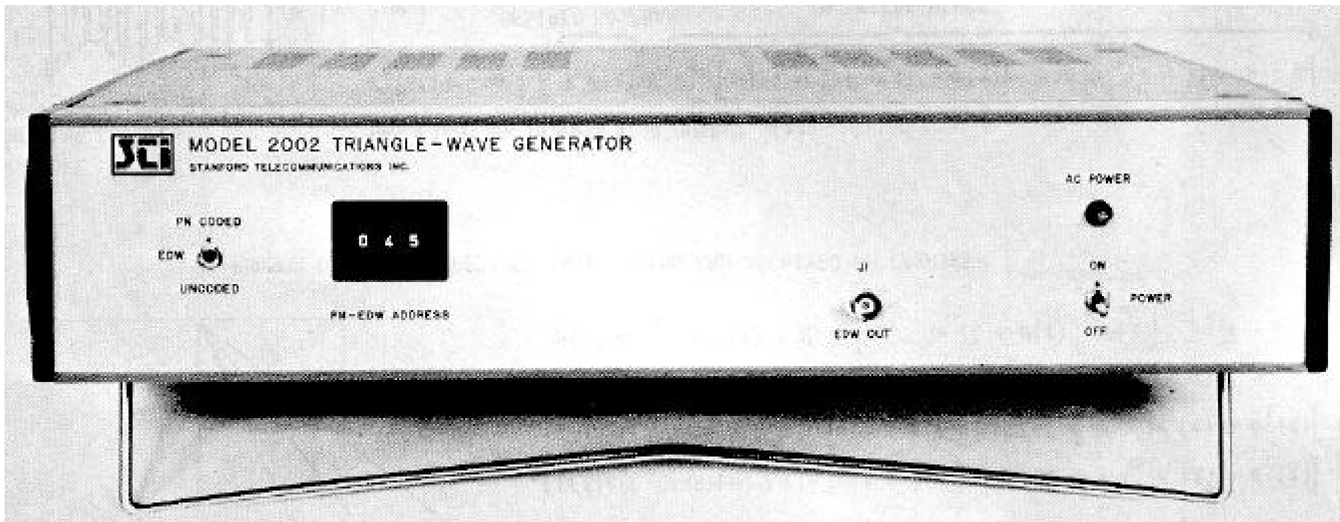
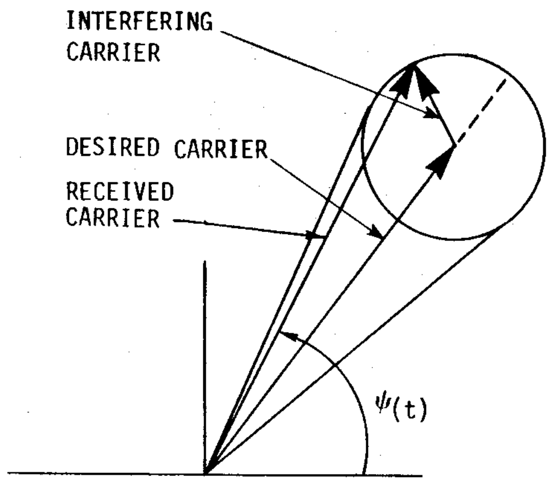



**FIGURE 4   STI 2002 CODED EDW GENERATOR**




$\beta^2$ = INTERFERENCE-TO-DESIRED CARRIER POWER RATIO

$\omega_0, \omega_i$ = RF CENTER FREQUENCIES

$x_0, x_i$ = BASEBAND FREQUENCY-MODULATION SIGNALS

$e_0, e_i$ = CODED EDW

$K_m$ = MODULATOR SENSITIVITY, Hz/VOLTS

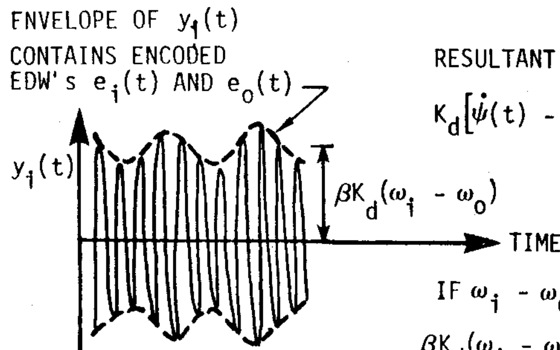
$K_d$ = DEMODULATOR SENSITIVITY, VOLTS/Hz

$$\begin{cases} \text{DESIRED CARRIER: } A \sin\left[\omega_0 t + \phi(t)\right], \text{ WHERE } \dot{\phi} = K_m\left[x_0(t) + e_0(t)\right] \\ \text{INTERFERENCE: } A\beta \sin\left[\omega_i t + \theta(t)\right], \text{ WHERE } \dot{\theta} = K_m\left[x_i(t) + e_i(t)\right] \end{cases}$$

RESULTANT RECEIVED SIGNAL PHASE FOR $\beta << 1$ (OR $I/C << 1$):

$$\psi(t) = \omega_0 t + \phi(t) + \beta \sin\left[(\omega_i - \omega_0)t + \theta(t) - \phi(t)\right]$$

a. PHASOR DIAGRAM OF DESIRED, INTERFERING, AND RECEIVED FM CARRIERS




RESULTANT FM DEMODULATOR OUTPUT $z(t)$:

$$K_d\left[\dot{\psi}(t) - \omega_0\right] = \underbrace{K_d\dot{\phi}(t)}_{\text{DESIRED BASEBAND MODULATION SIGNAL}} + \underbrace{K_d\beta\left[(\omega_i - \omega_0) + \dot{\theta}(t) - \dot{\phi}(t)\right] \sin\left[(\omega_i - \omega_0)t + \theta - \phi\right]}_{\text{FM INTERFERENCE } y_i(t)}$$

IF $\omega_i - \omega_0 >> \dot{\theta} - \dot{\phi}$ THEN ENVELOPE OF $y_i(t)$ IS $g(t) = \beta K_d\left[\omega_i - \omega_0 + \dot{\theta}(t) - \dot{\phi}(t)\right] =$

$$\beta K_d(\omega_i - \omega_0) + \beta K_d K_m\left[x_i(t) - x_0(t)\right] + \beta K_d K_m\left[e_i(t) - e_0(t)\right]$$

b. CONVOLUTIONAL INTERFERENCE COMPONENT $y_i(t)$ VS TIME AT FM DEMODULATOR OUTPUT

**FIGURE 5 EDW SIGNATURE RECOVERY CONCEPT**

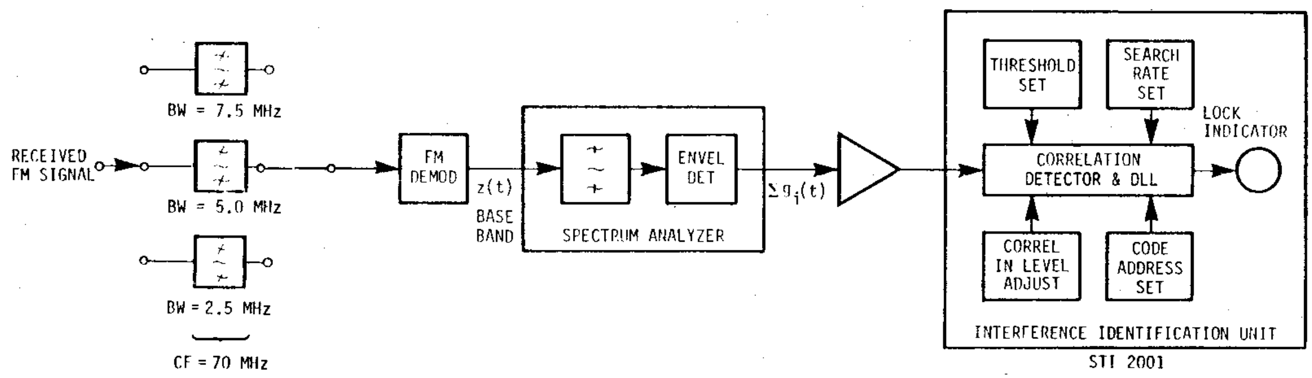



**FIGURE 6 LABORATORY TEST SYSTEM FOR TESTING THE STI 2001 INTERFERENCE IDENTIFICATION UNIT**

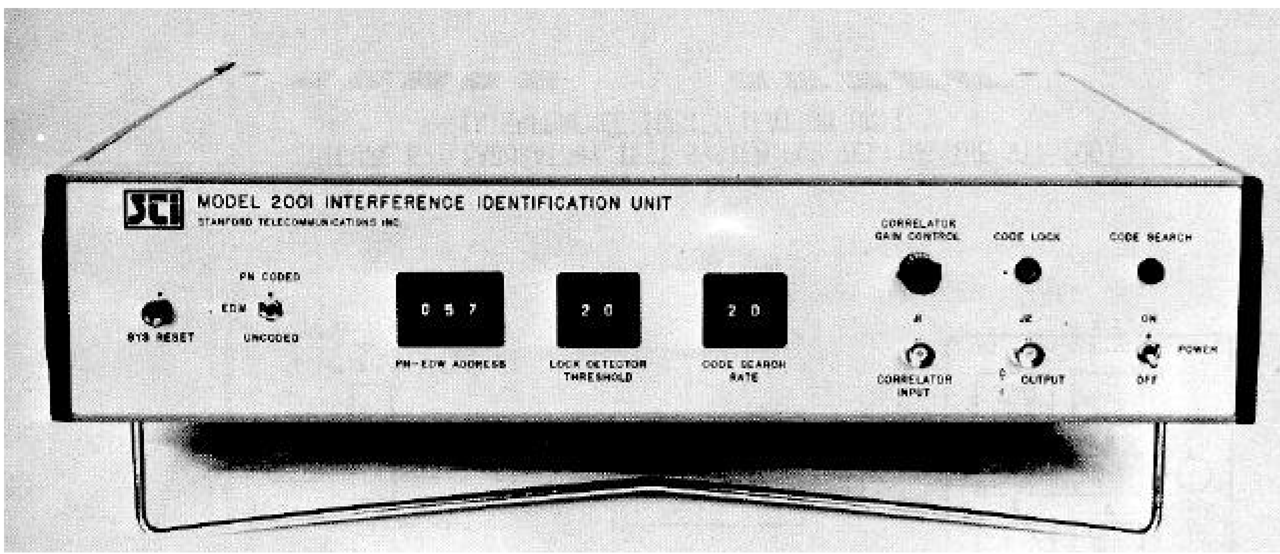



**FIGURE 7 STI 2001 INTERFERENCE IDENTIFICATION UNIT**

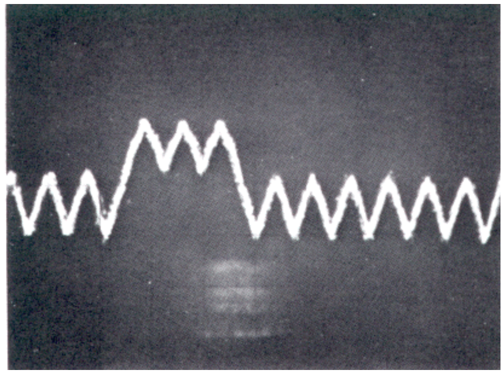


**FIGURE 8 TYPICAL RECOVERED EDW AT ENVELOPE DETECTOR OUTPUT PORT**

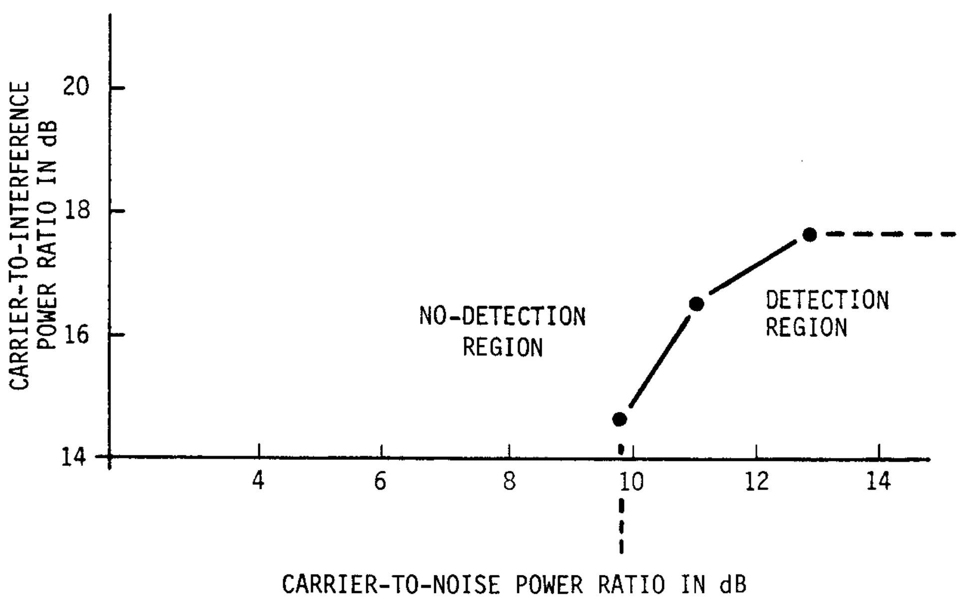



**FIGURE 9 TYPICAL LABORATORY RESULTS OF EDW SIGNATURE DETECTABILITY LIMITS**

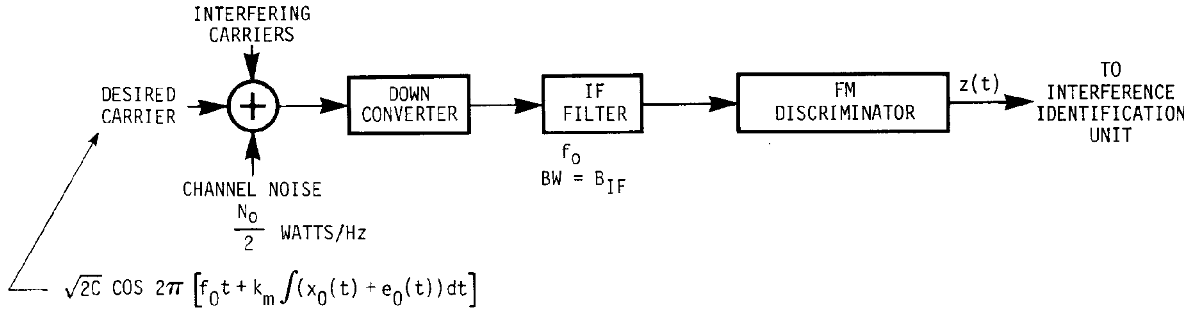
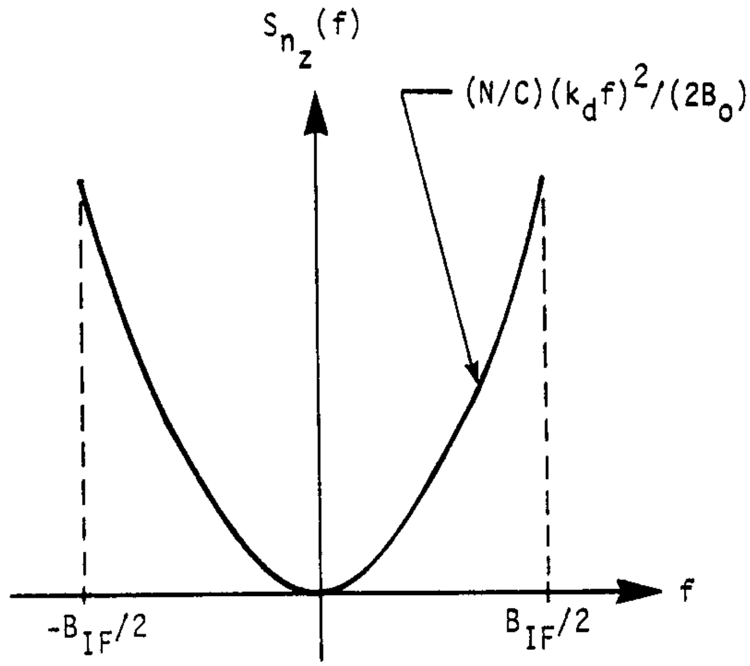



**FIGURE 1-A SIMPLIFIED FM RECEIVER BLOCK DIAGRAM**




C/N = DESIRED CARRIER-TO-NOISE POWER RATIO OVER THE OCCUPIED BW OF THE DESIRED CARRIER

$B_{IF}$ = PREDEMODULATION FILTER BW

$B_0$ = OCCUPIED BW OF THE DESIRED CARRIER

**FIGURE 2-A NOISE POWER SPECTRUM AT FM DISCRIMINATOR OUTPUT**

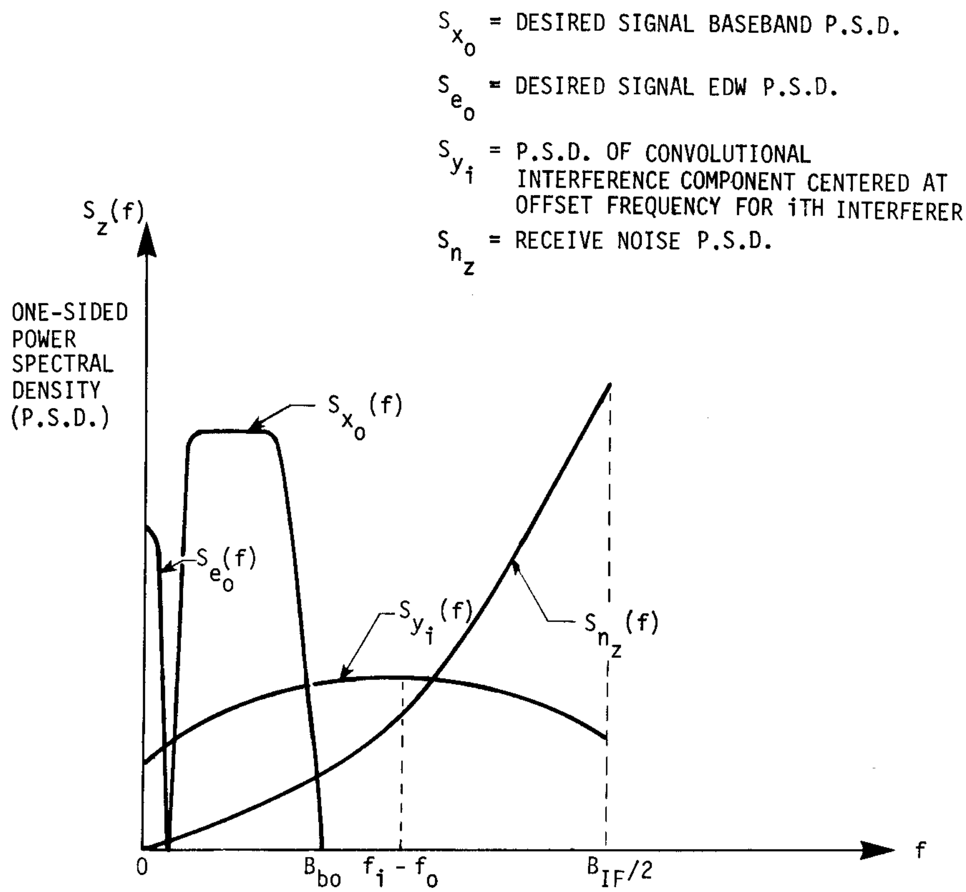



**FIGURE 3-A POWER SPECTRAL DENSITIES OF VARIOUS COMPONENT IN THE FM DEMODULES OUTPUT IN THE PRESENCE OF THE ITH INTERFERER**

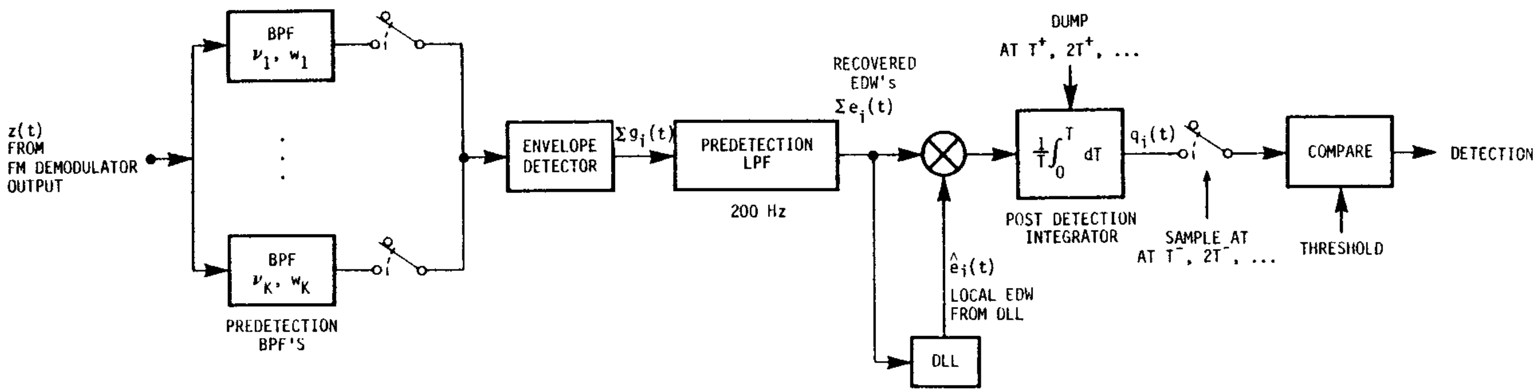



**FIGURE 4-A SIMPLIFIED BLOCK DIAGRAM OF THE INTERFERENCE IDENTIFICATION UNIT**