A Simple If In-Phase Combiner and Its Performance for Point-to-Point Radio Relay System with Space Diversity

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Abstract

The implementation of a simple analog in-phase combiner is presented for a high capacity radio relay system with space diversity. It provides good performance in terms of simple hardware and easy control, and measured results are in good agreement with simulated ones. To suggest practical applications, signatures with/without diversity are measured for STM-1 signal of 64-QAM radio relay system combined with a 13-tap equalizer, and they provided more than 25 dB fade depth at 10^{-3} BER under the frequency selective fading condition.

Key words : Microwave Radio Relay System, Space Diversity, Combiner, Signature, Frequency Selective Fading.

Ⅰ. Introduction

Modern radio relay systems have proved to be a highly reliable transmission medium operating in full compliance with ITU-T/ITU-R quality objectives[1], which mainly results from diversity techniques and signal processing of a powerful adaptive time and frequency domain equalizer[2]. The system performance is controlled not only by the signal to noise ratio, but also by the amount of dispersion in the channel transfer function. Channel dispersion introduces cross-talk between the orthogonal rails and inter-symbol interferences. It degrades a bit error ratio and sometimes causes unacceptable performance or outage[3]. Multipath fading is often accompanied by very severe channel dispersion because of the destructive interferences among the received multiple signals with different path delays[4].

As a countermeasure to the effects of multipath propagation, the space diversity is commonly used to obtain a second replica of the transmitted signal at the receiver[5],[6]. So it is very important to realize good algorithms for producing a combined signal from receiving antennas. Generally two algorithms, implemented by an analog circuit with some digital controls[7],[8], are mainly applied to even SDH radio relay system with 30 or 40 MHz channel spacing. One is a maximum amplitude power(MAP) combiner[7]. It is sure of the maximum power at one frequency within the in-band, usually fixed at the channel center, but can not assure a minimum in-band dispersion(MID). The other is the MID combiner[9]. It concerns only the minimum in-band dispersion, however it can not guarantee the maximum power throughout the in-band. Considering hardware complexity and controllability, the MAP combiner is relatively simple. So it has the advantage of narrow band system applications with less than 40 MHz channel spacing, and furthermore the modified MAP combiner working with weighted spectrum was also suggested for the wide band 2-carrier system with 80 MHz channel spacing[10].

In this paper, a simple IF in-phase combiner and its performance are discussed in view of hardware implementations and practical applications. By introducing a simple normalizing circuit in the phase detector, more improved results are obtained even for a severe notch depth. To show practical space diversity applications, numerical and experimental results including system signature are illustrated here for an STM-1 signal of 64-QAM radio relay system with a 13-tap adaptive equalizer[11].

Ⅱ. Formulation and Design

For the channel model of the space diversity, the multi-path fading is caused by destructive interference between rays arriving at a receiving antenna via different paths, such as the one taken by direct rays and the interfering rays refracted in the atmosphere layer or reflected off the earth's surface, as shown in Fig. 1.

The channel model adopted in digital radio relay system is mainly based upon the Rummler's 2-path model[4], and the received signal can be expressed by

$$R = a[1 - b\, e^{-j(\omega - \omega_n)\tau}] \qquad (1)$$

where a and b denote a flat loss term and a relative

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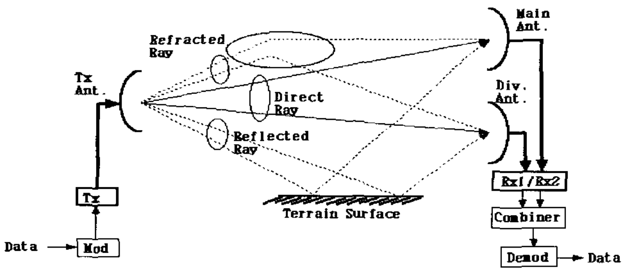


Fig. 1. Simplified space diversity transmission with combiner.

amplitude, respectively. τ is a relative time delay between direct and indirect rays, and ω_n means a notch angular frequency producing the minimum frequency response. The magnitude and the phase of Eq. (1) can be written as

$$|R| = a[1+b^2-2b\ \cos((\omega-\omega_n)\tau)]^{1/2} \tag{2}$$

$$\theta = \tan^{-1}[b\ \sin((\omega-\omega_n)\tau)/(1-b\ \cos((\omega-\omega_n)\tau))]. \tag{3}$$

The loss term A is expressed as $-20\log a$, and the relative notch depth B is given as $-20\log(1-b)$. Consequently the total fade depth is equal to $A+B$ at the minimum phase response for $\tau>0$ and $0<b<1$. The response is non-minimum phase when the sign of the delay is reversed($\tau<0$, $0<b<1$). The non-minimum phase state is also obtained when the relative amplitude of the delayed ray is greater than unity($\tau>0$, $b>1$), and the relative notch depth B is given as $-20\log(1-1/b)$.

From Eq. (1), the received signals from main and diversity antennas become

$$R_m = a_m[1-b_m\ e^{-j(\omega-\omega_m)\tau_m}] = |R_m|e^{j\theta_m} \tag{4}$$

$$R_d = a_d[1-b_d\ e^{-j(\omega-\omega_d)\tau_d}] = |R_d|e^{j\theta_d} \tag{5}$$

where subscript m and d denote the main and the diversity, respectively. Eqs. (4) and (5) can be easily calculated by Eqs. (2) and (3). Combining these two signals, the resultant output can be expressed in the form of

$$R_c = R_m + R_d\ e^{j\theta_c} \tag{6}$$

where θ_c denotes the phase difference between the received signals from main and diversity. In other words, it is the phase shift of the combiner. Then the combined signal power $P_c(w)$ is given by

$$P_c(w) = |R_c|^2. \tag{7}$$

Therefore the in-phase combiner operates the received signals such that two signals from main and diversity antennas are co-phased at the chosen frequency. Usually

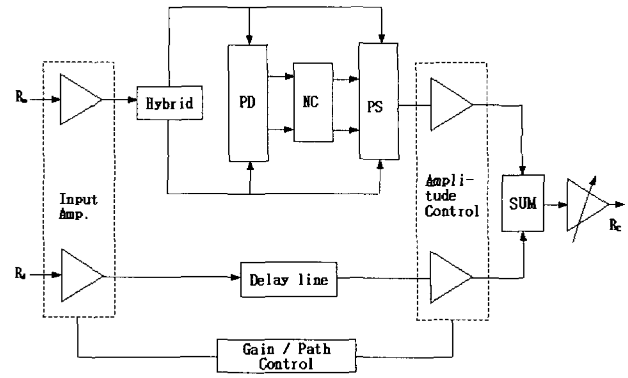


Fig. 2. Block diagram of in-phase combiner.

the combined power is maximum for the channel center frequency at the phase θ_c of the phase shifter. The combiner chooses a θ_c as follows

$$\begin{aligned}\theta_{\max.pwr} &= \theta_m - \theta_d \\ &= \tan^{-1}\left[\frac{b_m \sin((\omega-\omega_m)\tau_m)}{1-b_m\cos((\omega-w_m)\tau_m)}\right] \\ &\quad - \tan^{-1}\left[\frac{b_d \sin((w-w_d)\tau_d)}{1-b_d\cos((w-w_d)\tau_d)}\right].\end{aligned} \tag{8}$$

Now consider the design of the in-phase combiner as shown in Fig. 2. In general it mainly consists of 5 functions such as phase detector(PD), normalizing circuit (NC), phase shifter(PS), co-phase summation(SUM), delay line including gain and control part.

Fig. 3 depicts the block diagram of phase detector which yields a phase difference between two received signals. Using the quadrature hybrid circuit, the main signal can be divided into in- and quadrature-phase signals which are fed into multiplier circuit with the diversity signal, and the multiplier output generates two frequency components such as D.C. and the second harmonics of input signal. Therefore D.C. signal can be readily taken by the lowpass filter, which provides the phase difference of two received signals. To maintain a constant amplitude of the phase shifter output, it is necessary to introduce the normalizing circuit, followed by the phase detector. It means that the magnitude of

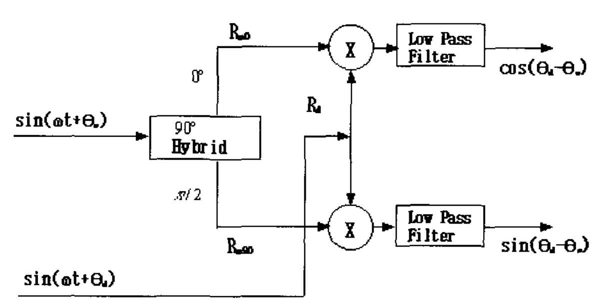


Fig. 3. Block diagram of phase detector.

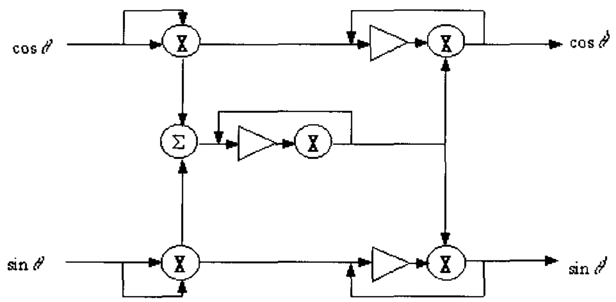


Fig. 4. Schematic of normalizing circuit.

vector signal of the phase detector output should be constant even for a severe fade. Fig. 4 illustrates a block diagram of the proposed normalization circuit, which consists of 5 analog multipliers and 2 differential amplifiers.

The phase shifter shown in Fig. 5 controls one of two phases, and then its output may be co-phased with respect to the other. The outputs of two variable gain amplifiers are given by

$$R_{o0} = R_{m0} \times a \cos(\theta_d - \theta_m) \quad (9)$$

$$R_{o90} = R_{m90} \times a \sin(\theta_d - \theta_m), \quad (10)$$

respectively, where a is constant. The output of the phase shifter is the sum of Eqs. (9) and (10), which is expressed by

$$R_{out} = a \sin(\omega_t + \theta_d). \quad (11)$$

In consequence, the phase θ_m of the main signal is shifted by the difference of $\theta_d - \theta_m$, which results in the same phase of the diversity signal.

In order to investigate the co-phase summation as well as the gain and path control, the co-phase summation, placed in the last part of Fig. 2, has a function of combining two signals with the same phase. Before combining two signals, to yield better performance of combiner, the gain and path control plays a role in adjusting signal gain appropriately and making a decision of path connection or not. If the signal difference between main and diversity is too large, there is no need for combining two signals because performance rather

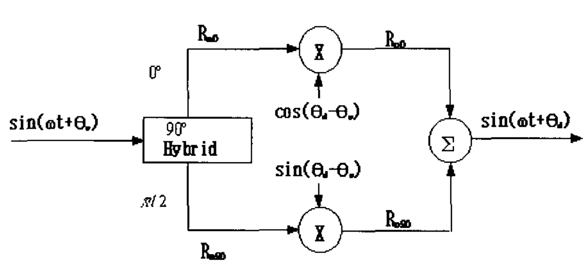


Fig. 5. Block diagram of phase shifter.

depends on a dominant signal. Therefore the cut-off condition on the signal path is conducted when the signal difference is greater than 17 dB in this study.

Also let's examine the delay line shown in Fig. 2. It is related with how to compensate the electrical length difference between two signals from main and diversity antennas. Although the combiner design is successfully carried out, the difference in electrical length causes degraded performance. In order to solve this, it is necessary to make the same length by means of

$$\delta l = \frac{5}{6} \frac{\delta\phi}{\delta f \sqrt{\varepsilon_r}} \quad (12)$$

where $\delta\phi$ is the locus of phase detector output on the oscilloscope in degrees, δf means the IF sweeping frequency of in-band in MHz, δl denotes the electrical length difference in meters, and ε_r is a relative dielectric constant of the compensating cable. If one can measure the arc of phase trajectory, $\delta\phi$, the length difference can be easily calculated, and then it should be added to the diversity path of the combiner. It is noted that if the compensation length is incorrect, a phase change of 360 degrees or more produces a continuous curve on the oscilloscope. However if well adjusted, then a small arc remains-ideally a dot.

Finally, prior to designing the in-phase combiner, Table 1 shows the required combiner specifications, which is suitable for STM-1 signal of 64-QAM radio relay system with 40 MHz channel. It is generally valid for multi-level QAM without any quality degradation. If other combiner schemes are introduced such as MID combiner or weighted MAP combiner, the parameters relevant to cut-off condition and combining gain listed in Table 1 may be changed due to RF bandwidth, con-

Table 1. Specification of designed in-phase combiner for transmitting STM-1 signal with 64-QAM modulation.

IF frequency / channel bandwidth	70 MHz / 40 MHz
Input / output power level	-10 dBm +/-1 dB
Input / output VSWR	<1.3
Input / output impedence	50 Ohm
Combining gain	>2.0 dB @BER: 10^{-3}
Amplitude flatness	<0.2 dB @BW: 32 MHz
Input / output return loss	<15 dB
Cut-off condition on signal path	>17 dB @Signal difference between main and diversity
Intermodulation distortion	>50 dBc

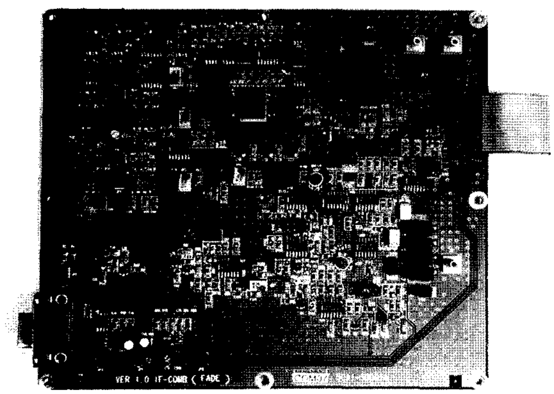


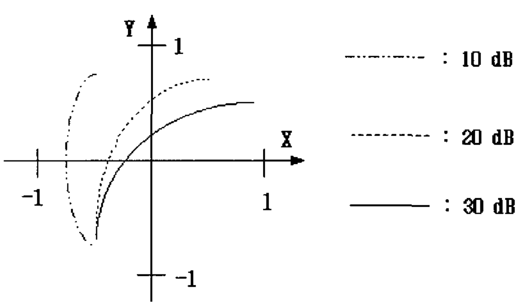
Fig. 6. Photograph of the implemented combiner.

trol algorithm and its implementation. Fig. 6 shows the photograph of implemented IF in-phase combiner, which has the 2-layer PCB with 15 cm×15 cm.

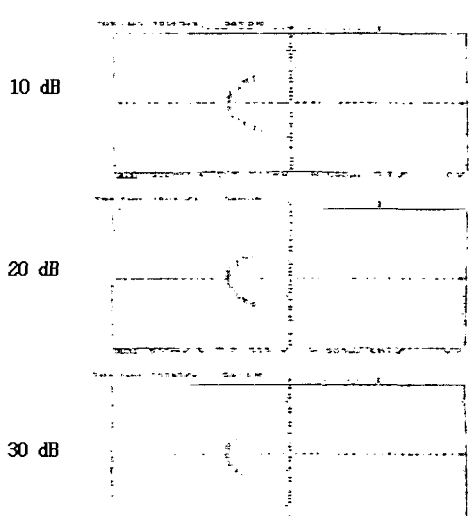
Ⅲ. Simulated and Experimental Results

It is assumed that the test inputs are STM-1 signals of 64-QAM with 40 MHz channel spacing in the frequency of 4.7 GHz. First, let's see the effect of how well the normalizing scheme is working even for a severe fading. For three notch depths of 10, 20, and 30 dB with 6.3 nsec time dely generated by the fading simulator, Fig. 7 shows the measured voltage vector loci of the phase detector output with/without normalizing circuit. The fading in the diversity occurred at the fixed frequency of 70 MHz and that in the main happened in the frequency of 50 to 90 MHz. The loci shown in Fig. 7(a) are not constant radius, but are getting worse as the notch depth increases. However those shown in Fig. 7(b) reveal a nearly constant radius of arc even for 30 dB although the arc length is gradually decreasing as the fading depth increases.

Second, consider simulated and experimental results for the implemented combiner as shown in Fig. 8. Fig. 8(a) shows the simulated result, where the notch frequencies of main and diversity are 61 and 66 MHz, respectively, with the same notch depth of 30 dB. Fig. 8(b) is the measured one. Since the combiner input is 64-QAM modulated signals with frequency selective fading, the spectra of main and diversity show the

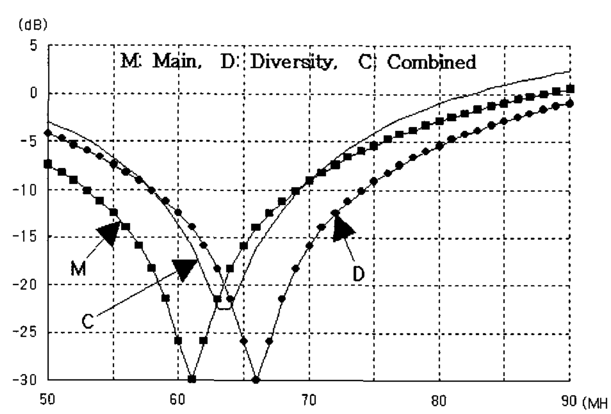


(a) Without normalization

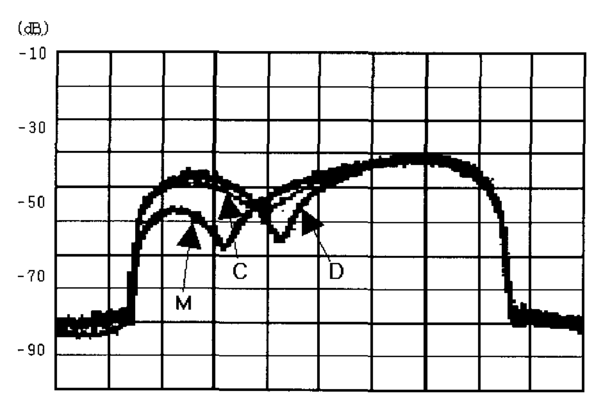


(b) With normalization

Fig. 7. Vector loci of phase detector output.

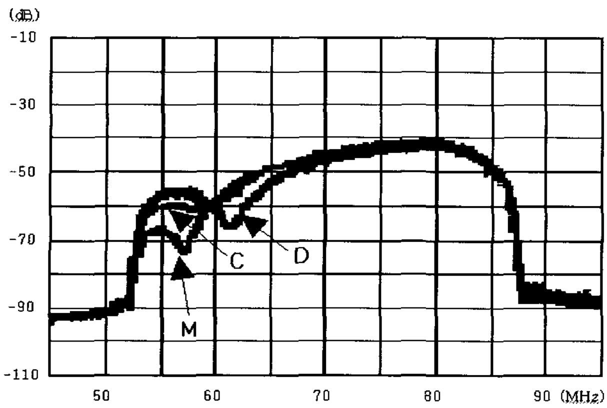


(a) Simulation

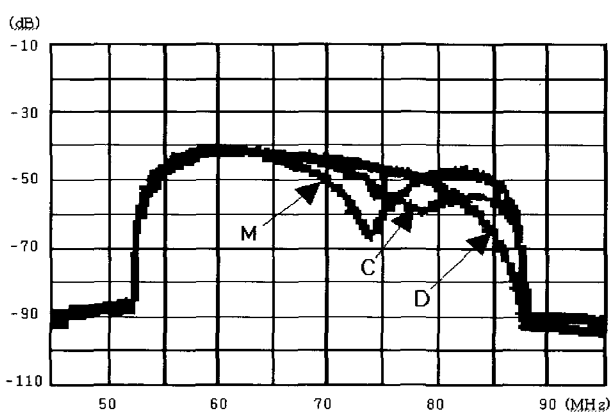


(b) Experiment

Fig. 8. Simulated and experimental results of combiner.



(a) 57 and 61 MHz



(b) 74 and 88 MHz

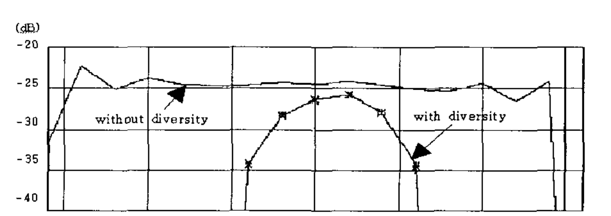
Fig. 9. Experimental results of combiner for two notch frequency sets.

different depth and shape due to the location of notch frequency. However the overall behaviors of both combined results are in good agreement, and the minimum dispersion value of the combined spectrum differs from about 1.0 dB regarding that of the spectrum with less notch depth, namely, diversity.

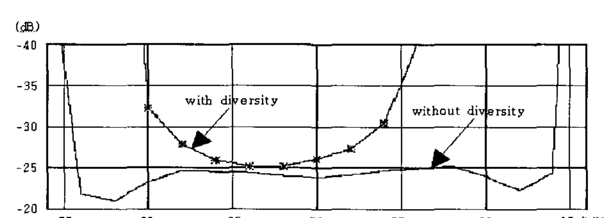
As another example, Fig. 9 displays the measured results according to two sets of notch frequency, 57 and 61 MHz in Fig. 9(a) as well as 74 and 88 MHz in Fig. 9(b). Since there are some differences between notch frequencies corresponding to main and diversity for each case, the combined result and its shape may be different. The dispersion of the combined spectrum directly depends upon the location of the notch frequency and its depth. Even if notch depth is very deep upto 30 dB, the degree of dispersion is greatly improved. However it still remains and may cause the degraded system performance. In order to overcome this, it is well known that the adaptive equalizers in the time and the frequency domains are widely used and can remarkably counteract the distorted spectrum[2]. From all results it is noted that the minimum dispersion of the combined spectrum is improved by about 7.5 dB in comparison to that of the spectrum with less notch depth of main and diversity.

On the other hand, let's to evaluate system performance working with the combiner including the analog slope equalizer and the 13-tap time domain equalizer[11]. For 4.7 GHz radio relay system at a given 10^{-3} BER, signatures are measured for minimum and non-minimum phase fades, respectively. Fig. 10(a) shows two curves for the minimum phase fade with/ without space diversity. The notch in the diversity is fixed at 70 MHz with 22 dB depth, and that in the main is varied according to the frequency and its depth. In the similar way, signatures for the non-minimum phase fade can be obtained as shown in Fig. 10(b). As can be seen from two results, the width of signature curve without diversity is not usually much larger than the occupied channel bandwidth of about 35 MHz, reflecting the fact that the most damaging notches are ones that occur in-band.

Finally, in order to check signatures with diversity working with the combiner, Fig. 10(a) shows better performance compared with Fig. 10(b). This implies that the implemented IF in-phase combiner is better behavior for the minimum phase fade in view of H/W operation. In consequency the outage probability of system is a function of the dispersive fade margin depending upon the signature curve and its depth[3]. So the internal area of signature with diversity, which means greater than 10^{-3} BER, is much smaller than that without diversity. This makes it possible to reduce the outage time by a factor of the diversity improvement, and hence the availability can be significantly improved by using the space diversity technique[5].



(a) Minimum phase fade



(b) Non-minimum phase fade

Fig. 10. Measured system signature.

Ⅳ. Conclusions

In this paper, the implementation of a simple analog in-phase combiner has been examined for STM-1 signal of 64-QAM radio relay system. The measured results of combiner have been in good agreement with simulated ones based upon channel modeling, and the difference is less than about 1.0 dB. In order to keep the stable output of phase detector, adoption of the simple normalizing circuit provides more improved results even for a severe fading upto 30 dB by enlarging the voltage range of phase detector output.

In addition, to evaluate system performance for diversity working with the 13-tap equalizer, signatures for minimum and non-minimum phase fades were measured and provided good results by more than 25 dB fade depth at 10^{-3} BER. It has shown that since the outage area of signature with diversity is remarkably reduced in comparison to that of signature without diversity, more improvement on availability could be obtained. Therefore it is expected that the implemented combiner is applicable to the space diversity for modern multi-level QAM system with less than 40 MHz channel spacing.

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